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Clayton R. Paul

University of Kentucky

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13 ABSTRACT

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This report examines the pertinent standards and reviews existing efforts to predict intrasystem compatibility via computer aided techniques.

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FOREWORD

The Post-Doctoral Program, under the direction of Dr. W. W. Everett, Jr., is a cooperative venture between Syracuse University (Department of Electrical Engineering), The U. S. Air Force Academy (Department of Electrical Engineering), Cornell University (School of Electrical Engineering), Georgia Institute of Technology (School of Electrical Engineering), Clarkson College of Technology (Department of Electrical Engineering), State University of New York at Buffalo (Department of Electrical Engineering), University of Kentucky (Department of Electrical Engineering), Purdue University (School of Electrical Engineering), the University of Adelaide (Department of Electrical Engineering) in South Australia and the Rome Air Development Center. The Post-Doctoral Program provides, via contract, the opportunity for faculty and visiting faculty at the participating universities to spend a year full-time on exploratory development efforts with the post-doctorals splitting their time between RADC and the educational institutions.

The author of this report is Dr. Clayton Paul. He received the B.S.E.E. degree from the The Citadel (1963), the M.S.E.E. degree from Georgia Institute of Technology (1964), and the Ph.D. degree from Purdue University (1970).

He was on the Georgia Tech faculty for a year in 1964-1965 and joined Purdue University as a graduate instructor in September 1965. His other professional experience was with American Telephone and Telegraph and with the Dupont Savannah River Project. His research interests are in network theory and linear multivariable systems. He is a member of Sigma Epsilon Pi and Tau Beta Pi.

This report has been reviewed by the Office of Information (OI) and is releasable to the National Technical Information Service (NTIS).

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Samuel D. Zaccari

Approved: SAMUEL D. ZACCARI
Chief, Compatibility Branch
Reliability and Compatibility Division

Joseph J. Naresky

Approved: JOSEPH J. NARESKY, Chief
Reliability and Compatibility Division

FOR THE COMMANDER: Fred I. Diamond

FRED I. DIAMOND
Acting Chief, Plans Office

ABSTRACT

In the procurement of electronic systems by the Department of Defense, electromagnetic compatibility has generally been considered in the design phase by the application of somewhat rigid and rather general specifications to the individual equipments/subsystems.

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TABLE OF CONTENTS

I. INTRODUCTION

In the procurement of electronic systems by the Department of Defense, EMC has generally been considered in the design phase by the application of somewhat rigid and rather general specifications to the individual equipments/subsystems. For example, MIL-STD-461 places limits on the allowable emissions from a subsystem (a "black box") and also places limits on the susceptibility of subsystems to emissions from other subsystems. This MIL-STD basically attempts to insure a "quiet system," i.e., the specification controls undesired or noise type signals within the system.

The problem here is that the MIL-STD's (e.g., MIL-STD-461, MIL-STD-826, MIL-STD-6181) are not written individually for each system under consideration. They are a set of specifications which are to be applied to all subsystems/equipments the Air Force (as well as all departments of DOD for the current MIL-STD-461) purchases. The present subsystem EMC specification, MIL-STD-461, seems to have evolved over the past 30 years and it is difficult to surface any rationale for the limits in these specifications. In fact, it seems that it is fundamentally impossible to attach any rationale to these general specifications at all. To do so, one would have to assume that one set of specifications could insure compatibility for such a wide class of applications.

To illustrate the deficiencies in the present approach, consider the following. The electromagnetic energy emitted by subsystems will be coupled to other subsystems through various transfer paths. For example, signals on a power line which are not at the power frequency may be undesirable

since they may cause interference by coupling to other subsystems as a result of direct conduction through the common power supply, radiation to nearby sensitive receiving antennas or radiation to other wires and equipment cases. Also the signal on the line at the power frequency will couple to other subsystems, possibly causing interference. What would be very desirable to know are 1) the levels of the desired as well as undesired emissions, 2) the portions of these emissions which are incident upon each susceptible device due to transferral through each coupling path and 3) the quantitative susceptibility of devices to the incident energy. If one can determine these items then a meaningful determination of compatibility can be made.

Herein lies the primary deficiency in the present approach to EMC of applying the same EMC subsystem specifications to many different systems. Simply placing limits on allowable emissions and susceptibilities of subsystems will not necessarily insure compatible operation of these subsystems. In certain cases, the actual emissions from the subsystems (even though they do not exceed the limits set by the MIL-STD's) when transferred through the various coupling paths may well exceed the prescribed susceptibility limit (set by the MIL-STD's) of a device. On the other hand, the limits may be too stringent requiring unnecessary suppression measures. Therefore, to achieve compatibility for a system and optimize the required interference suppression measures, one should determine the energy coupled between the subsystems.

Clearly a subsystem will be affected to some degree by all sources of energy (e.g., desired and undesired, wires, cases and antennas, etc.). A fundamental requirement for insuring compatibility of a system then is the

consideration of all paths of electromagnetic energy transfer. Not only must one consider the totality of these paths but the total amount of electromagnetic energy transferred must be determined; at least the "significant" energy must be determined. Obviously, the terms "all paths" and "total amount of transferred electromagnetic energy" are only conceptual topics. In a practical application, achieving these objectives would be an impossible task. We are only emphasizing the obvious point that all sources of energy will to some degree affect each subsystem in an actual deployment. This comment points out that not only should one consider coupling of energy generated within the system but also energy generated outside the system which is incident upon the subsystems should not be disregarded. Also one must consider desired or functional signals as well as undesired or nonfunctional signals. For example, one cannot neglect the power supply fundamental frequency (e.g., 60HZ, 400HZ) in determining subsystem compatibility since the subsystems will be affected to some degree in an actual deployment by these signals.

A system can be defined in many ways and is subject to a great deal of variability depending upon the particular needs. One may define a resistor as a "system." On the other hand, one may define the entire universe as a "system." Between these two extremes lies a compromise definition of a system. For example one may define an entire precision approach radar complex as a system. Within this system may be included the facilities at the PAR site to communicate with the "outside" world such as the VHF voice communications.

Clearly the definition of a system depends upon the situation at hand. Within a system we may isolate subsystems. The definition or delineation

of these subsystems generally is a compromise of two considerations. If one breaks the system into many subsystems which are physically small, then the analysis may become too unwieldy. For example, one could define the subsystems for the PAR complex at the component level, i.e., resistors, inductors, etc. However, if one defines the subsystems on a more macroscopic "black box" level, e.g., transmitters, power supplies, etc., then the resulting "model" of the system consisting of an "interconnection" of these black boxes may not admit an acceptable degree of accuracy in the system analysis. For example, one may model a communications receiver as a frequency dependent sensitivity curve. Yet this macroscopic model does not predict other effects such as intermodulation, crossmodulation, etc. Alternately, the receiver could be modeled at the component level to include all these effects with the resulting complex analysis. One could conceive of an electronic system as a set of "black boxes" which are connected by wires (cables) and antennas. In other words, these boxes would communicate or interact with each other through the transfer of electromagnetic energy via coupling between wires, antennas and the boxes themselves.

In order to determine quantitatively the coupling of electromagnetic energy between the various subsystems, a mathematical system model should be determined. This model would consist of analytical models for generators and receptors (or processors) of electromagnetic energy along with analytical models of the various transfer paths. The use of the term system model does not imply that there will be one model for all systems. Obviously this cannot be the case. A system model is merely a mathematical interconnection of the various subsystem mathematical models to represent the actual system.

In a realistic situation, the subsystems are affected by electromagnetic energy from outside the system as well as from inside the system which will be denoted as intersystem effects and intrasystem effects respectively. Intersystem effects result from natural as well as man made sources. Examples of natural sources are cosmic or extraterrestrial sources and thunderstorms. The man made sources include desired as well as undesired or extraneous energy. The radiation from neon signs and spurious emissions from a transmitter are examples of undesired man made sources, whereas the fundamental frequency emission of a transmitter is an example of a desired man made source. It seems reasonable to assume that intrasystem effects are the more difficult to consider if we assume that the physical proximity of systems is much greater than that of the subsystems within a particular system. In view of this assumption, one may model intersystem effects as point sources or uniform field sources in many cases. Additionally, in determining intersystem coupling between two distinct systems, the wire-to-wire coupling between systems seems intuitively to be negligible in comparison with intrasystem wire-to-wire coupling. Also antenna-to-antenna coupling for intrasystem analysis will require near field considerations in many cases whereas far field considerations are generally sufficient for intersystem considerations.

In the past, intrasystem coupling of electromagnetic energy generated within the system has been considered through the adjustment of the general subsystem specifications by granting waivers. This adjustment of subsystem specifications is based upon engineering experience and results in many cases in overdesign with associated unnecessary costs and in some cases in system incompatibilities necessitating costly retrofit. These inaccuracies should not be totally attributed to the project engineers. The physical size and

complexity of modern systems present a monumental problem of interaction which in many cases exceeds human analytical capability.

Pearlston has pointed out many inconsistencies in the primary subsystem EMC specifications; MIL-STD-461, 462, 6181, and 326[1]. The specifications basically limit the amount of undesired radiated and conducted electromagnetic energy from equipments (e.g. cable conduction, box radiation, etc.) and provide upper bounds on the susceptibility of a device to radiated and conducted electromagnetic energy which enters at various "ports" of the device (e.g. wires, antennas, etc.).

The obvious problem, however, is not the subsystem specifications but that the totality of the coupling paths is not presently considered in sufficient analytical detail. For example, one could model the transfer paths (e.g., wire-to-wire, [illegible]-to-wire, antenna-to-wire, etc.) with mathematical transfer functions which accurately reflect the physical location and proximity of the subsystems within the system and the types of coupling paths (e.g., conducted, radiated, wire-to-wire, etc.). The inputs to these mathematical transfer functions could be the limits in the above applicable MIL-STD's for the generated extraneous (undesired) energy (we are assuming as a worst case that the devices actually emit signals at the levels of the limits) and design determined levels for desired energy. To consider the total electromagnetic energy within a system, one must include all desired sources of electromagnetic energy (e.g., transmitter fundamental) as well as all undesired sources (e.g., transmitter spurious frequencies). The total effect of these sources on each susceptible device or receptor could be determined through an in-depth quantitative analysis based on the mathematical description of the various coupling paths. The total electromagnetic effect at the receptor port could be compared with the susceptibility limit for the appropriate MIL-STD to determine

if system compatibility will be achieved by subsystem compliance to these specifications. The point to be made here is that if all sources of electromagnetic energy (desired and undesired) are considered in determining the electromagnetic energy incident upon a subsystem or device and if all sources of extraneous energy conform to the above applicable MIL-STD's then the system will be guaranteed compatible (within the confidence of the models) if the total energy incident upon each subsystem (determined analytically above) does not "violate" the applicable susceptibility specification in the case of undesired receptors or a design determined susceptibility limit for desired receptors. This of course assumes that all devices and subsystems will be properly tested to insure their compliance with the appropriate MIL-STD's. It should be clear that if an in depth quantitative analysis such as the above is not performed, then little confidence can be placed in system compatibility from compliance to the MIL-STD's.

The above approach has difficulties however. In order to insure compatibility one must consider all sources of electromagnetic energy, undesired and desired, intrasystem and intersystem. This will be impossible to implement even on the largest computers for even moderate size systems. One then must make assumptions (hopefully reasonable ones) to made the computation tractable. The mathematical modeling of the generators, coupling paths and receptors can be simple or difficult depending upon the effort one wishes to place in this area. The modeling of the receptors or processors of electromagnetic energy seems to be the more difficult of the three. The processing of electromagnetic energy by these receptors is in many cases quite subjective and does not admit a simple description. For example the level of noise which produces interference in a communications receiver is highly dependent upon the listener (i.e., a novice listener or a "seasoned veteran").

A common (and quite reasonable) assumption in the determination of the coupling descriptions are the assumptions of linearity and time-invariance of the coupling mediums. By linearity of the coupling medium we mean the following. Consider two sources of electromagnetic energy G_1 and G_2 and one receptor, R_1. Basically, linearity of the coupling paths means that if G_1 produces effect E_1 incident upon R_1 with G_2 "turned off" and G_2 produces effect E_2 incident upon R_1 with G_1 "turned off" then G_1 and G_2 operating simultaneously will produce an effect $E_1 + E_2$ incident upon R_1. Clearly this is not a rigorous definition of linearity but should serve our purposes here.

Basically the assumption of time-invariance of the coupling paths means that the mathematical parameters sufficiently describing the coupling paths do not change with time. These assumptions of linearity and time-invariance of the coupling paths merely simplify the mathematical analysis and have been implicitly assumed almost universally (and with good justification) in the field of EMC. With these assumptions we may determine frequency domain transfer functions for the coupling paths since a frequency domain description of a linear, time-invariant system is equivalent to a time domain description.

The inputs to the transfer functions can be frequency domain descriptions of the generator outputs. For example, the time domain output of a generator can be represented equivalently in the frequency domain through the Fourier Transform. The result of transferring the generator outputs through the transfer functions of the coupling paths yield a frequency domain description of the effects of the generators on the receptor which if desired may be converted to a time domain description through the inverse Fourier Transform. At this point the modeling difficulties increase greatly. What is required is a mathematical description of the effect of this incident energy upon the receptor.

This is clearly very subjective and difficult to provide correctly in most caces.

For example, one may wish to determine a frequency dependent susceptibility threshold for the receptor. A common way of doing with is to apply sinusoidal input signals A $\sin\omega t$ to the input port and at each frequency $f = \omega/2\pi$ increase A until a "malfunction" occurs. Then if the total incident effect due to all the generators at the receptor port exceeds this threshold at a particular frequency then surely one can say that the device is "susceptible." However, if the incident effect (described in the frequency domain) at a particular frequency does not exceed the threshold at any frequency then can one say that the device is not susceptible? It is doubtful that one can be assured of this for the following reason. If the receptor processes the incident "effects" linearly then we may represent this as a linear system. If an input u(t) is applied to a linear, time-invariant system with impulse response h(t), then the output, y(t), will be given by the convolution

$$y(t) = \int_{-\infty}^{\infty} h(t-\tau)\, u(\tau)\, d\tau \tag{1-1}$$

If the Fourier Transform of u(t), $U(\omega)$, is obtained, then the output "spectrum" will be given by

$$Y(\omega) = H(\omega)\, U(\omega) \tag{1-2}$$

where $H(\omega)$ is the Fourier Transform of h(t). The inverse Fourier Transform of $Y(\omega)$ yields y(t) as

$$y(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} H(\omega)\, U(\omega)\, e^{j\omega t}\, d\omega \tag{1-3}$$

and the output is a <u>summation</u> of the effects of each component of $U(\omega)$. Therefore the susceptibility level determined by applying sinusoids to the device in the above manner may not be the relevant indicator of susceptibility due to this

summation of effects even for receptors which process the incident effects linearly. This idea of a frequency dependent susceptibility threshold has been in constant use in the field of EMC and clearly needs some justification. Some general attempt is made in the MIL-STD's to account for this effect through the use of a broadband specification. However, it seems that the reason for instituting this type of specification was the realization that measurement equipments do not have infinitesimally small bandwidth.

To simplify the analysis, one might also make the assumption of "unilateral effects." By this we mean that the receptors do not "load" the generators and one may simply determine a "transfer function" between each generator and each receptor whose output is dependent on the input yet the input is not affected by the output. Also one might assume that classes of generators do not interact. For example, we might assume that the output of a transmitter is not affected by the wire voltages. If these assumptions are made, then the computational difficulties are reduced substantially since one may determine through independent analyses the effects at a receptor due to each source individually. Then however, <u>all</u> these effects must be added to yield the <u>total</u> effect of all sources on the receptor. This last point is very important if one wishes to discuss compatibility. Considering partial effects of sources is not a valid way of determining compatibility.

Another difficult area in this type of modeling effort would be the determination of intersystem effects. However, as pointed out above, these effects could be modeled on a worst case basis much more easily than intrasystem effects.

The main deficiency in the present approach to system EMC is that the energy coupled between subsystems is not presently considered in sufficient analytical detail. Also, intersystem effects are not considered except through adjustment of the limits when intersystem sources are known (e.g., when a high power radar

is in close proximity to the system). We should also point out that the MIL-STD's generally exclude functional signals from consideration (e.g., the 60 HZ power frequency). However, even if compliance to the above MIL-STD's was shown to insure system compatibility through an indepth analysis as suggested above, there would be no optimization in the design of interference suppression since the MIL-STD's are formulated apriori, e.g., without consideration of the needs of the specific system.

This is not to say that there is no need for general type subsystems specifications as are presently in existence. There are many cases in which the final system configuration is either unknown or not fixed. Also the system environment may not be known or fixed as with an aircraft. However, more meaningful general type specifications could be developed for classes of general systems through the long term quantitative experience gained in analyzing particular types of these systems through an indepth quantitative analysis based on the evaluation of the various coupling paths within each system.

A comprehensive survey of the problems and a proposed solution in the determination of intrasystem compatibility can be found in [2].

In conclusion, we see that the application of subsystem limits in their present useage, e.g., MIL-STD-461, is adequate in many cases for the following reasons.

1) Simply placing limits on subsystems will not insure compatibility of these subsystems unless the various coupling paths are considered in detail, analytically.

2) Subsystem compliance to the specifications generally will not result in optimum system design from the standpoint of required interference suppression, minimum system weight and complexity, etc.

3) The impact of granting waivers on the overall system compatibility cannot be easily or accurately assessed.

4) The impact of future design changes, i.e., system modifications in terms of subsystem additions or deletions, cannot be easily or accurately assessed.

5) The impact of design trade-offs on system compatibility cannot be easily or accurately assessed.

6) The expense of system tests generally exceeds that which is required since critical areas cannot be determined which would allow focusing of the test effort.

7) Subsystem/equipment test data cannot be related to system performance.

The quantitative analysis suggested above would provide a way of determining more precisely the cost and mission effectiveness of a system.

II. TECHNICAL INTRODUCTION

Here we will discuss the concepts of mathematically modeling the transfer or coupling paths. We will assume that the transfer paths consist of radiation and direct conduction (for wires only) between wires, antennas and boxes. We then are faced with modeling the coupling modes of wire-to-wire, wire-to-antenna, wire-to-box, antenna-to-wire, antenna-to-antenna, antenna-to-box, box-to-wire, box-to-antenna, and box-to-box.

2.1 Wire-to-wire

We will begin the discussion by reviewing basic transmission line theory. These results and derivations are given in many texts[3-11]. Consider the parallel pair of wires (e.g., a transmission line) of length $\mathcal{L}$ and spacing between centers D shown in Figure 1. We can model the transmission line as in Figure 2 where ℓ_1, r_1, ℓ_2, r_2 are the self inductance and resistance per unit length for wires 1 and 2 respectively, c,g are the capacitance and conductance per unit length between the wires and m is the mutual inductance between the wires per unit length.

The voltage and current along the line are functions of two variables x and t where x represents distance along the line and t represents time, i.e., v(x,t), i(x,t). Therefore we may write using elementary lumped circuit analysis techniques for a small, infinitesimal length of line, Δx

$$v(x + \Delta x,\ t) - v(x,t) = -\ell\Delta x\,\frac{\partial\, i(x,t)}{\partial t} - r\Delta x\, i(x,t)$$
$$i(x + \Delta x,\ t) - i(x,t) = -c\Delta x\,\frac{\partial v(x + \Delta x, t)}{\partial t} - g\,\Delta x\, v(x + \Delta x,\ t) \qquad (2\text{-}1)$$

where the current in line 1, $i_1(x,t)$, equals the current in line 2, $i_2(x,t)$,

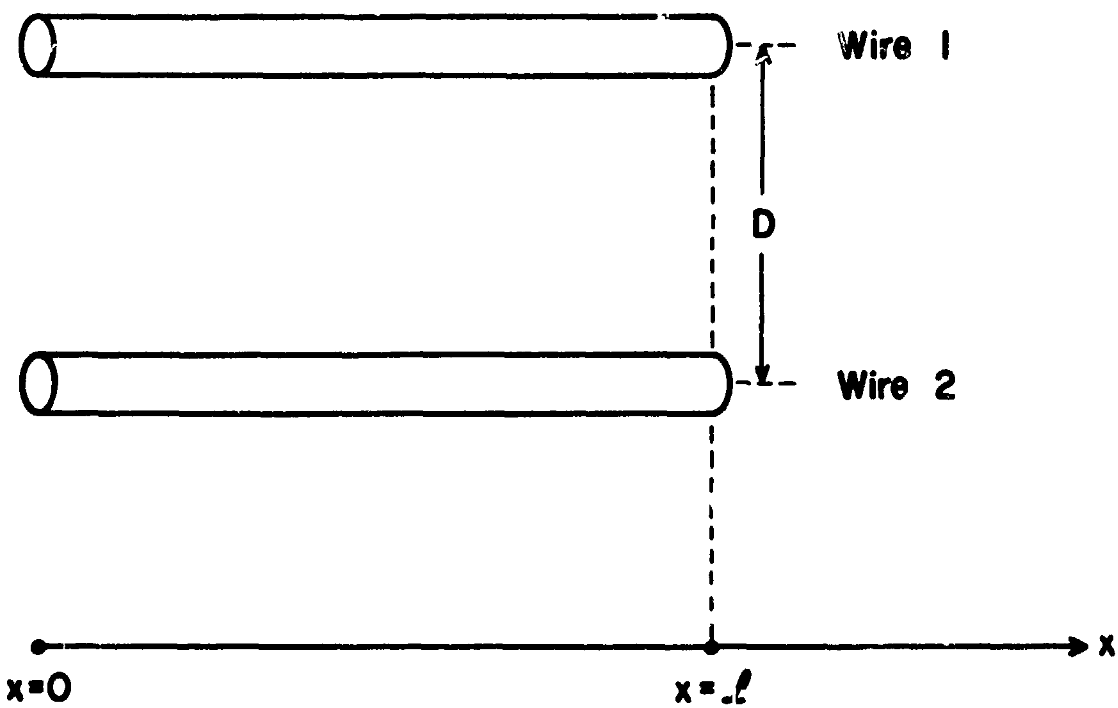


FIGURE 1

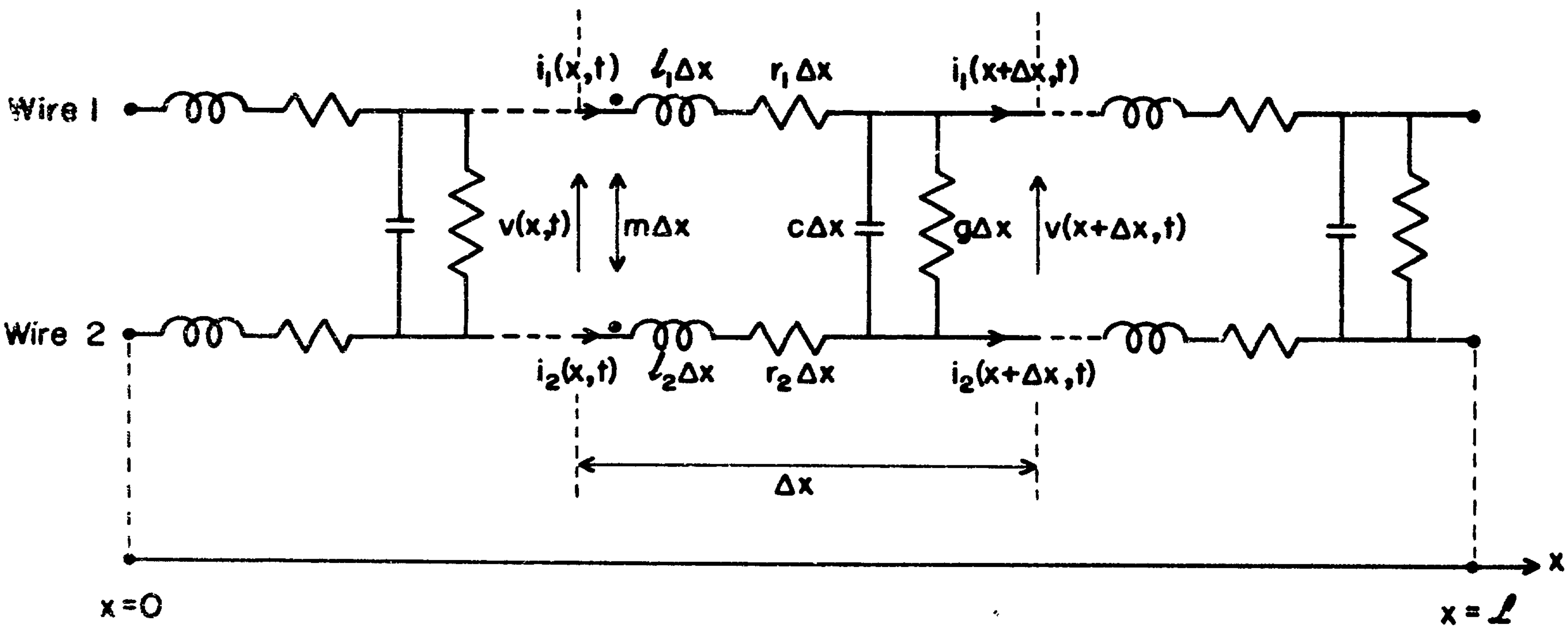

Wire 1
Wire 2
$i_1(x,t)$
$\ell_1\Delta x$
$r_1\Delta x$
$i_1(x+\Delta x,t)$
$v(x,t)$
$m\Delta x$
$c\Delta x$
$g\Delta x$
$v(x+\Delta x,t)$
$i_2(x,t)$
$\ell_2\Delta x$
$r_2\Delta x$
$i_2(x+\Delta x,t)$
Δx
x
x = 0
$x = \ell$

FIGURE 2

i.e., $i_1(x,t) = -i_2(x,t) = i(x,t)$ (and $\ell = (\ell_1 + \ell_2 + 2m)$ and $r = (r_1 + r_2)$).

Dividing (2-1) by Δx and taking the limit as $\Delta x \to 0$ yields the standard partial differential equations describing the transmission line

$$\frac{\partial\, v(x,t)}{\partial x} = -\ell\, \frac{\partial i(x,t)}{\partial t} - r\; i(x,t)$$

$$\frac{\partial\, i(x,t)}{\partial x} = -c\, \frac{\partial v(x,t)}{\partial t} - g\; v(x,t) \qquad (2\text{-}2)$$

Now if we assume all variables on the line are sinusoidal of frequency $f = \omega/2\pi$ then we may write without loss of generality

$$v(x,t) = V(x)e^{j\omega t}$$

$$i(x,t) = I(x)e^{j\omega t} \qquad (2\text{-}3)$$

where V(x) and I(x) are functions only of x (this amounts to the usual separation of variables method of solving partial differential equations).

Substituting (2-3) into (2-2) yields

$$\left(\frac{dV(x)}{dx}\right) e^{j\omega t} = -j\omega\ell\; I(x)e^{j\omega t} - r\; I(x)e^{j\omega t}$$

$$\left(\frac{dI(x)}{dx}\right) e^{j\omega t} = -j\omega c\; V(x)e^{j\omega t} - g\; V(x)e^{j\omega t} \qquad (2\text{-}4)$$

By cancelling $e^{j\omega t}$ we have

$$\frac{dV(x)}{dx} = -\,(j\omega\ell + r)\; I(x) = -\, z\; I(x)$$

$$\frac{dI(x)}{dx} = -\,(j\omega c + g)\; V(x) = -\, y\; V(x) \qquad (2\text{-}5)$$

a set of ordinary differential equations. Differentiating (2-5) yields

$$\frac{d^2V(x)}{dx^2} = -\, z\, \frac{dI(x)}{dx} = zy\; V(x) \qquad (2\text{-}6)$$

and similarly

$$\frac{d^2I(x)}{dx^2} = zy\; I(x) \qquad (2\text{-}7)$$

These equations are solved from ($D = \frac{d}{dt}$ operator).

$$(D^2 - zy)\, V(x) = 0 \tag{2-8}$$

There are two roots of the characteristic equation $(D^2 - zy) = 0$, $\pm\sqrt{zy}$, and the solution is

$$\begin{aligned} V(x) &= Ae^{\gamma x} + Be^{-\gamma x} \\ I(x) &= Ce^{\gamma x} + De^{-\gamma x} \end{aligned} \tag{2-9}$$

where $\gamma = \sqrt{zy}$.

Differentiating the second equation of (2-9) yields

$$\frac{dI(x)}{dx} = \gamma Ce^{\gamma x} - \gamma De^{-\gamma x} = -y\, V(x) \tag{2-10}$$

from (2-5). Therefore the equations of the line become

$$\begin{aligned} V(x) &= Ae^{\gamma x} + Be^{-\gamma x} \\ I(x) &= -\frac{A}{z_0} e^{\gamma x} + \frac{B}{z_0} e^{-\gamma x} \end{aligned} \tag{2-11}$$

where $z_0 = \sqrt{z/y}$ is the characteristic impedance of the line.

Now to solve (2-11) we must apply boundary conditions to obtain A and B. Suppose the terminations for the line are as shown in Figure 3 where Z_S and Z_L are complex impedances and $v_s(t) = V_S e^{j\omega t}$ is a complex sinusoid. Then we have from Figure 3 $V(0) = V_S - I(0)Z_S$, $V(\ell) = I(\ell)Z_L$. Therefore we may solve the following equations for A and B

$$\begin{aligned} V(0) &= A + B = V_S - I(0)Z_S \\ V(\ell) &= Ae^{\gamma\ell} + Be^{-\gamma\ell} = I(\ell)Z_L \\ I(0) &= \frac{A}{z_0} - \frac{B}{z_0} \\ I(\ell) &= \frac{A}{z_0} e^{\gamma\ell} - \frac{B}{z_0} e^{-\gamma\ell} \end{aligned} \tag{2-12}$$

We then may obtain $v(x,t) = V(x)e^{j\omega t}$ and $i(x,t) = I(x)e^{j\omega t}$ for any point x on the line and any sinusoidal voltage source $v_s(t) = V_S e^{j\omega t}$.

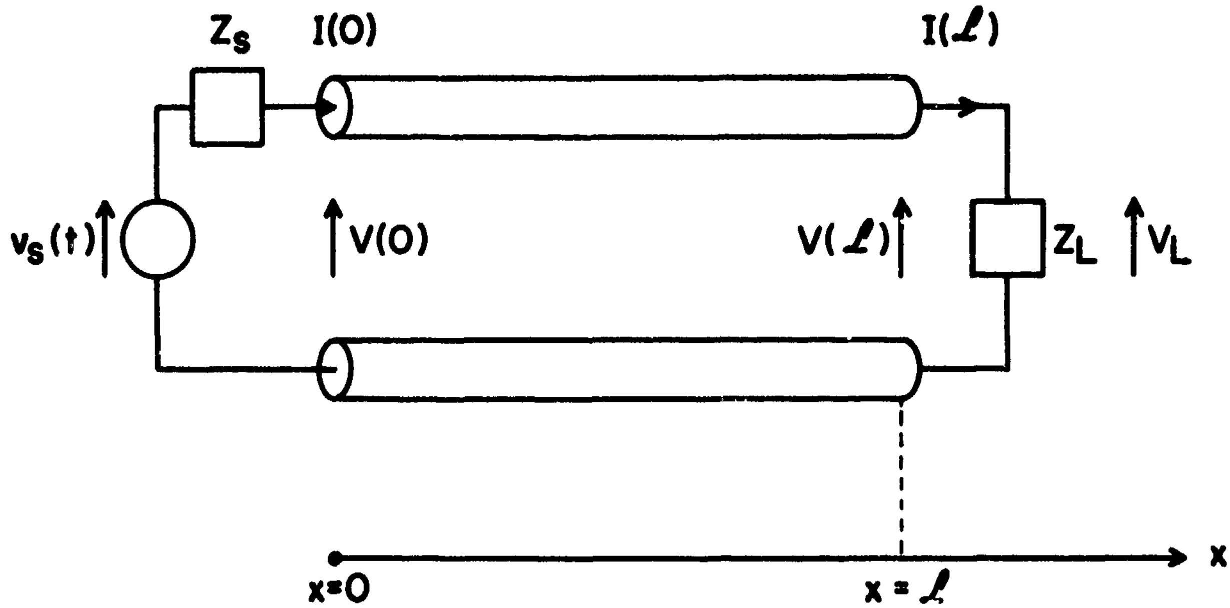


FIGURE 3

We could also produce a set of equations relating I(0), V(0), I(ℓ) and V(ℓ), the terminal variables of the line, in two port parameter form as

$$V(0) = V(\ell)\cosh\gamma\ell + I(\ell)z_0\sinh\gamma\ell$$
$$I(0) = I(\ell)\cosh\gamma\ell + \frac{V(\ell)}{z_0}\sinh\gamma\ell \qquad (2\text{-}13)$$

Now the difficult part is computing the capacitance, inductance, resistance and conductance per unit length for the lines. To illustrate typical calculations for these quantities we will consider a pair of infinitely long conductors as in Figure 4. The wires have radii $r_{\omega1}$ and $r_{\omega2}$. Let us assume that wire 1 and wire 2 carry q_1 and q_2 coulombs of charge per unit length distributed uniformly over the surface. Consider a single wire in free space with q coulombs per unit length as in Figure 5. The electric field due to this charge distribution will extend radially from the conductor since the charge is assumed to be uniformly distributed over the surface of the wire. From Gauss's Law (which also is one of Maxwell's equations), "The surface integral of the normal component of the electric flux density, $\vec{D}$, over any closed surface equals the charge enclosed, Q," i.e.,

$$\iint \vec{D}\cdot\vec{da} = Q \qquad (2\text{-}14)$$

where $\vec{da}$ is a unit surface area normal to $\vec{D}$. If we assume that the medium is homogeneous and isotropic then $\vec{D} = \epsilon\vec{E}$ where ϵ is the permittivity of the medium, then (2-14) becomes

$$\iint \vec{E}\cdot\vec{da} = Q/\epsilon \qquad (2\text{-}15)$$

Therefore using Gauss's Law where the area is a "pill box" with the wire passing through its center, of R radius and length 1 meter,

$$Q = q \times (1\text{ meter}) = \iint \vec{D}\cdot\vec{da} = \epsilon\vec{E}\,2\pi R \qquad (2\text{-}16)$$

since the surface area is $2\pi R \times 1$. Therefore, the magnitude of $\vec{E}$, $|\vec{E}|$, is

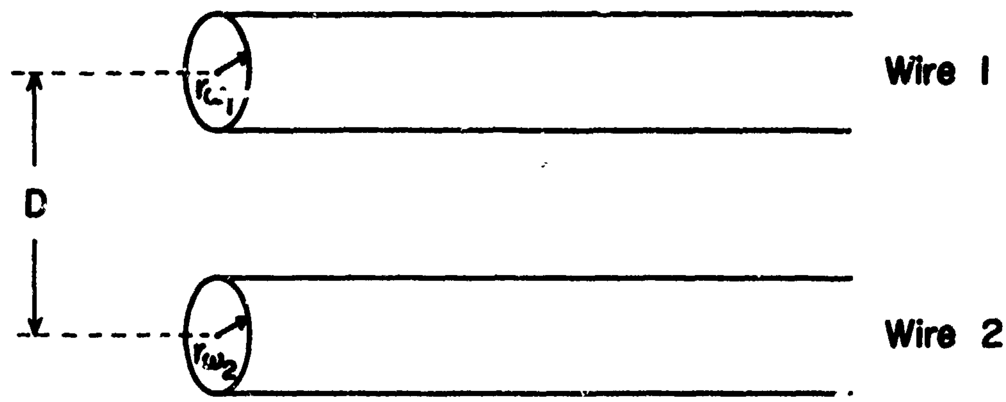


FIGURE 4

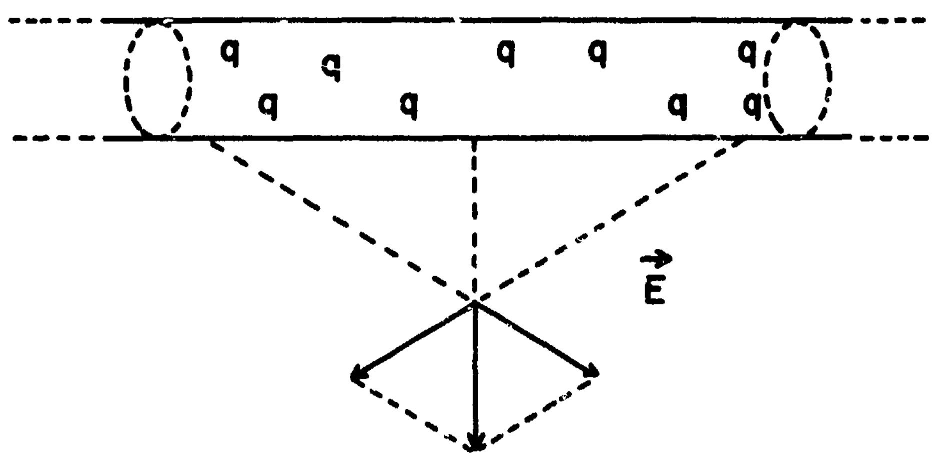


FIGURE 5

given by

$$|\vec{E}| = \frac{q}{2\pi R\epsilon} \tag{2-17}$$

and is directed radially from the wire.

Now looking "end-on" on the wire we will calculate the potential difference between two points P_1 and P_2 as in Figure 6. In calculating the potential difference between P_1 and P_2 we use the definition of potential difference

$$v_{12} = \int_{D_1}^{D_2} \vec{E} \cdot \vec{d\ell} \tag{2-18}$$

where $\vec{d\ell}$ is the unit path length. Along ℓ_2 the integral is zero since this is an equipotential surface (due to the fact that $E \cdot d\ell = |E|\ |d\ell| \cos\theta$ and θ along this path is 90^o). Along path ℓ_1 we simply have

$$\begin{aligned} v_{12} &= \int_{D_1}^{D_2} \frac{q}{2\pi\epsilon\ell}\, d\ell \\ &= \frac{q}{2\pi\epsilon} \ln \frac{D_2}{D_1} \text{ volts} \end{aligned} \tag{2-19}$$

The capacitance between two bodies is defined as the ratio of charge on the two bodies to the potential difference between the two bodies, i.e.,

$$C = Q/V \tag{2-20}$$

The capacitance per unit length of the two wire line is therefore

$$c = q/v_{12} \tag{2-21}$$

where q is the sum of the charge per unit length for each line and v_{12} is the potential difference between the conductors.

Returning to the original problem of the determining the capacitance between the two wires, let us assume the "medium" is linear and use superposition. Assume $q_2 = 0$. Therefore wire 2 being a conductor is an equipotential surface and the potential difference due to q_1 is from our previous derivation

$$\frac{q_1}{2\pi\epsilon} \ln \frac{D}{r_{\omega 1}} \tag{2-22}$$

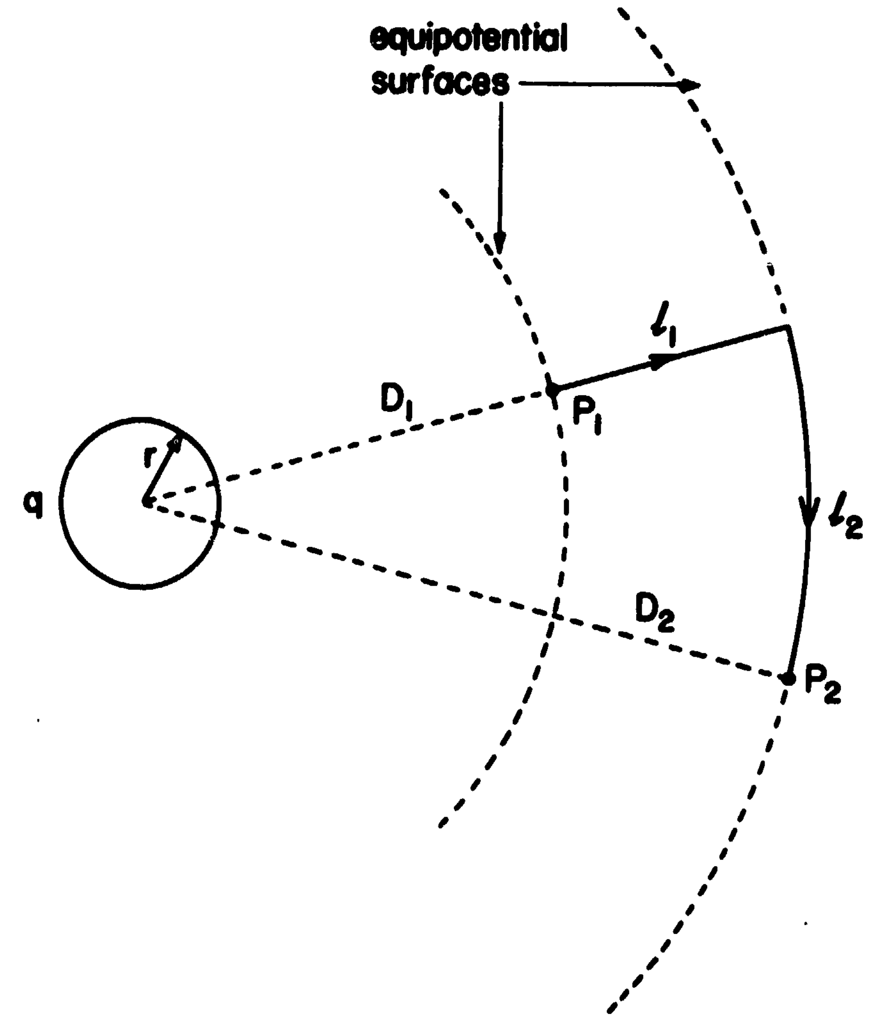


FIGURE 6

where we use the path of integration shown in Figure 7, and ℓ_2 and ℓ_3 are equipotential surfaces.

Now we find v_{12} due to q_2 with $q_1 = 0$ as

$$-\frac{q_2}{2\pi\epsilon} \ln \frac{D}{r_{\omega 2}} = \frac{q_2}{2\pi\epsilon} \ln \frac{r_{\omega 2}}{D} \tag{2-23}$$

Therefore the total potential difference is

$$v_{12} = \frac{q}{2\pi\epsilon} \left\{ \ln \frac{D}{r_{\omega 1}} + \ln \frac{r_{\omega 2}}{D} \right\} \tag{2-24}$$

where we assume $q_1 = -q_2 = q$. Rewriting (2-24)

$$v_{12} = \frac{q}{2\pi\epsilon} \ln \left(\frac{D^2}{r_{\omega 1} r_{\omega 2}} \right) \tag{2-25}$$

and therefore

$$c = \frac{q}{v_{12}} = \frac{2\pi\epsilon}{\ln \left(\frac{D^2}{r_{\omega 1} r_{\omega 2}} \right)} \tag{2-26}$$

Now we compute the inductance per unit length, ℓ. Consider a single wire in a medium of permeability μ. We assume a current I is uniformly distributed throughout the wire. Inductance ℓ (per unit length) is defined as ψ/I where ψ is the total amount of flux (per unit length) linking (enclosing in a complete path) the current I. So, ψ is what we are seeking. First, we determine ψ_{int}, the flux internal to the wire, and then we determine ψ_{ext}, the flux external to the wire.

Consider Ampere's Law

$$\int_{\substack{\text{closed} \\ \text{path}}} \vec{H} \cdot \vec{d\ell} = I \text{ (current enclosed by path)} \tag{2-27}$$

Now the $\vec{H}$ field in the wire is totally tangential (assumed infinitely long wire and uniform current distribution), therefore

$$\int \vec{H} \cdot \vec{d\ell} = 2\pi r\, H_T = I_r \tag{2-28}$$

where H_T is the tangential magnetic field and I_r is the portion of I enclosed

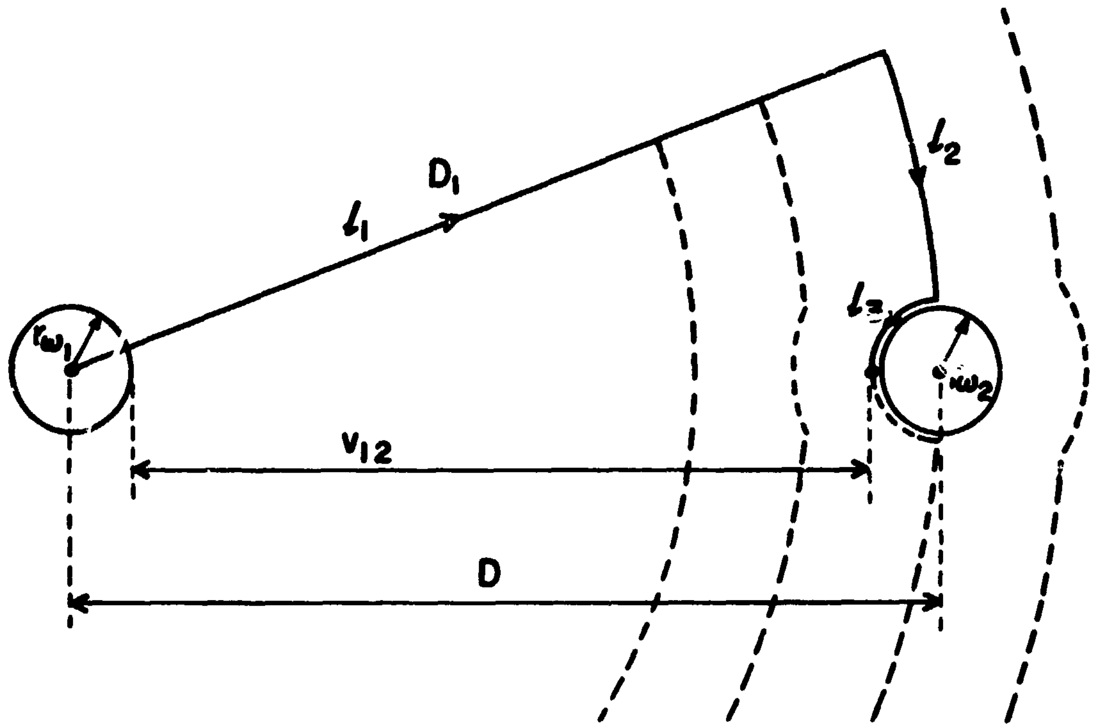


FIGURE 7

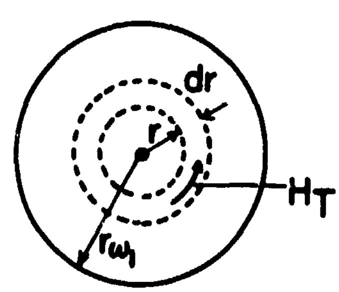


FIGURE 8

in radius r as shown in Figure 8. If we assume the current is uniformly distributed in the wire, then the portion of I enclosed in radius r is given by

$$I_r = \frac{(\pi r^2)}{(\pi r_{\omega 1}^2)} I \tag{2-29}$$

Therefore

$$2\pi r\, H_T = \frac{\pi r^2}{\pi r_{\omega 1}^2} I \tag{2-30}$$

or

$$H_T = \frac{r}{2\pi r_{\omega 1}^2} I \tag{2-31}$$

and the magnetic flux density (per unit length) is

$$B_T = \mu_\omega H_T = \frac{\mu_\omega r_{\omega 1}}{2\pi r_{\omega 1}^2} I \tag{2-32}$$

where μ_ω is the permeability of the wire. In a tubular element of thickness dr and unit length along the line, the total magnetic flux (per unit length) is the product of B_T and the cross sectional area of the element normal to the flux lines or $\Phi = \int\int \vec{B} \cdot \vec{da}$ ($\vec{da}$ is the unit surface area) or

$$d\Phi = \frac{\mu_\omega r}{2\pi r_{\omega 1}^2} I \;\; dr \tag{2-33}$$

The flux linkage $d\psi$ per unit length due to the flux in the tubular element is given by the product of the flux per unit length and the fraction of the current linked. Thus

$$d\psi = \frac{\pi r^2}{\pi r_{\omega 1}^2} \quad d\Phi = \frac{\mu_\omega I r^3}{2\pi r_{\omega 1}^4}\, dr \tag{2-34}$$

Integrating from the center of the conductor to its outside edge to find the total flux linkages inside the conductor,

$$\begin{aligned} \psi_{int} &= \int_C^{r_{\omega 1}} \frac{\mu_\omega I r^3}{2\pi r_{\omega 1}^4}\, dr \\ &= \frac{\mu_\omega I}{8\pi} \end{aligned} \tag{2-35}$$

Now we must compute the flux linkages from flux external to the conductor. The H field is (tangential to the conductor) and

$$H_T = \frac{I}{2\pi r}$$
$$B_T = \frac{\mu I}{2\pi r} \tag{2-36}$$

where the permeability of the medium surrounding the wire is μ. The flux in the tubular element of thickness dr is ($\Phi = \int\int B \cdot da$) (see Figure 9)

$$d\Phi = \frac{\mu I}{2\pi r}\,dr \tag{2-37}$$

The total flux linkages between P_1 and P_2 then is given by

$$\Psi_{ext} = \int_{D_1}^{D_2} \frac{\mu I}{2\pi r}\,dr = \frac{\mu I}{2\pi} \ln \frac{D_2}{D_1} \tag{2-38}$$

Now consider the two wire line in Figure 10 with associated currents I_1 and I_2 directed into the paper. Let us determine the flux linkages for wire 1 with $I_2 = 0$ between the wire and point X. From equation (2-35) and (2-38) we have

$$\Psi_{11} = \frac{\mu_\omega I_1}{8\pi} + \frac{\mu I_1}{2\pi} \ln \frac{D_{1X}}{r_{\omega 1}} \tag{2-39}$$

Now for $I_2 \neq 0$ and $I_1 = 0$ we have

$$\Psi_{12} = \frac{\mu I_2}{2\pi} \ln \frac{D_{2X}}{} \tag{2-40}$$

Therefore the total flux up to point X linking I_1 is given by

$$\begin{aligned}\Psi_1 &= \Psi_{11} + \Psi_{12}\\ &= \frac{\mu_\omega I_1}{8\pi} + \frac{\mu I_1}{2\pi} \ln \frac{D_{1X}}{r_{\omega 1}} + \frac{\mu I_2}{2\pi} \ln \frac{D_{2X}}{D}\end{aligned} \tag{2-41}$$

Equation (2-41) may be rewritten as

$$\Psi_1 = \frac{\mu_\omega I_1}{8\pi} + \frac{\mu I_1}{2\pi} \ln \frac{1}{r_{\omega 1}} + \frac{\mu I_2}{2\pi} \ln \frac{1}{D} + \frac{\mu}{2\pi}\left[I_1 \ln D_{1X} + I_2 \ln D_{2X}\right] \tag{2-42}$$

Since $I_2 = -I_1$ (we may assume a closed circuit) we may write

$$\Psi_1 = \frac{\mu_\omega I_1}{8\pi} + \frac{\mu I_1}{2\pi} \ln \frac{1}{r_{\omega 1}} + \frac{\mu I_2}{2\pi} \ln \frac{1}{D} + \frac{\mu}{2\pi}\left[I_1 \ln \frac{D_{1X}}{D_{2X}}\right] \tag{2-43}$$

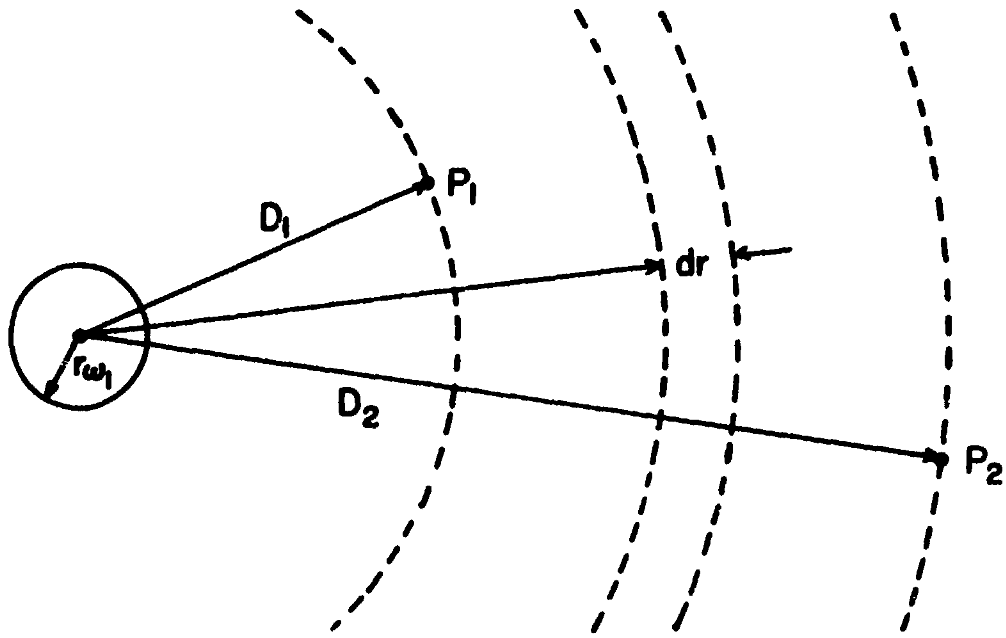


FIGURE 9

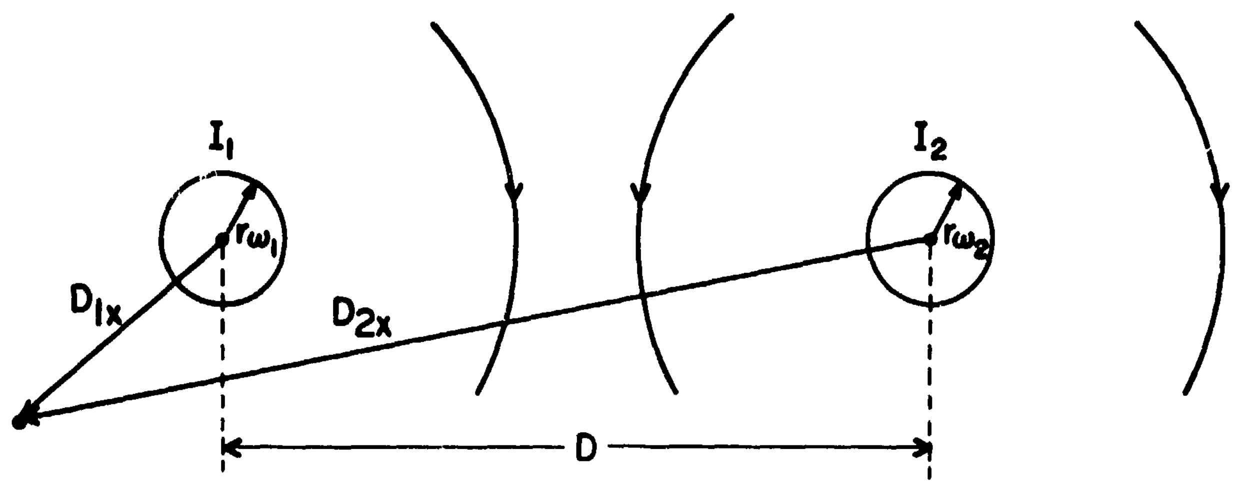


FIGURE 10

Allowing point X to go to infinity to determine the total flux linking conductor 1 due to I_1 and I_2 means that ${}^{D_{1X}}/_{D_{2X}} \rightarrow 1$ and the last term in (2-43) vanishes. Therefore we may interpret the self inductance of wire 1 as

$$\ell_1 = \frac{\mu_\omega}{8\pi} + \frac{\mu}{2\pi} \ln \frac{1}{r_{\omega 1}} \tag{2-44}$$

and the mutual inductance as

$$m_{21} = \frac{\mu}{2\pi} \ln \frac{1}{D} \tag{2-45}$$

Similarly for the second wire we have

$$\ell_2 = \frac{\mu_\omega}{8\pi} + \frac{\mu}{2\pi} \ln \frac{1}{r_{\omega 2}}$$

$$m_{12} = \frac{\mu}{2\pi} \ln \frac{1}{D} \tag{2-46}$$

Define $m = m_{12} = m_{21}$.

We may also derive the per unit length capacitance to ground for a wire suspended above a ground plane as shown in Figure 11. To solve this problem we make use of the method of images. Essentially this method states that we may effectively replace the ground plane with the image of the conductor as in Figure 12. From equation (2-26) we obtain

$$c_g = \frac{2\pi\varepsilon}{\ln \frac{4h^2}{r_\omega^2}} = \frac{\pi\varepsilon}{\ln \frac{2h}{r_\omega}} \tag{2-47}$$

Similarly one may determine the self inductance of a wire above a ground plane through the method of images. Also the mutual inductance and capacitance between two wires above a ground plane may be determined through the method of images.

The resistance per unit length, r, can be written as

$$r = r_{d.c.} + r_{a.c.}(f) \tag{2-48}$$

where $r_{d.c.}$ is the bulk resistance of the wire and $r_{a.c.}(f)$ is a frequency dependent resistance reflecting the fact that for higher frequencies, the

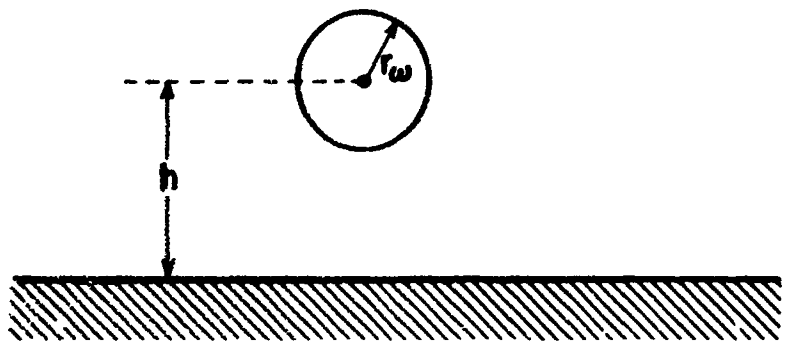


FIGURE 11

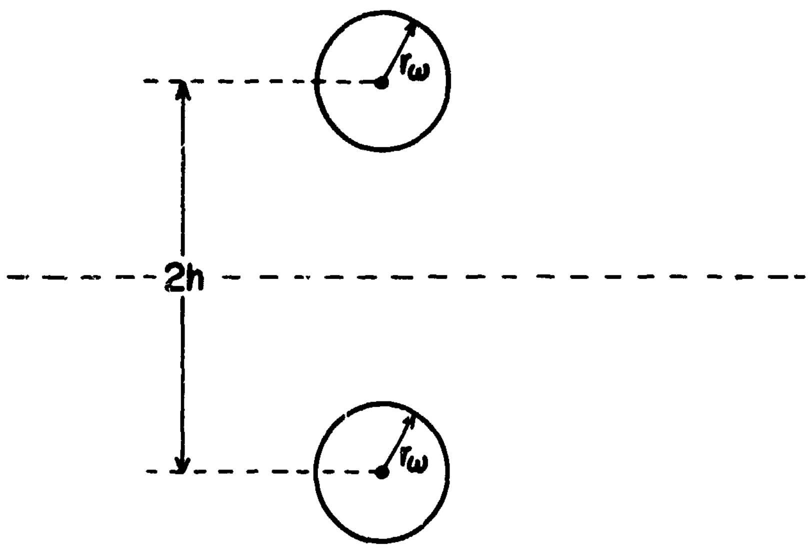
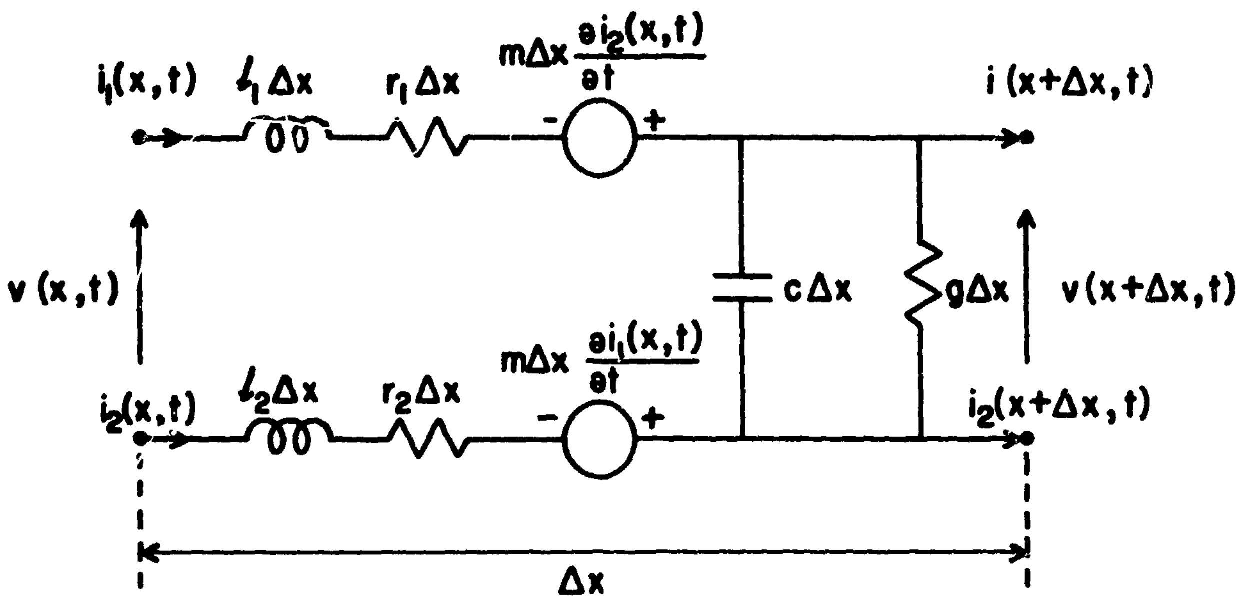


FIGURE 12

current tends to concentrate more closely to the outer surface of the conductor, i.e., skin effect. The above per unit length line parameters are derived in many texts.[3-8] Thus we may draw the equivalent circuit for an infinitesimal length of line ΔX as in Figure 13.

The derivation and solution of the partial differential equations of the line are quite straightforward. The difficult portion of the analysis is the derivation from elementary field theory of the per unit length parameters for actual lines (finite length, shielding, grounding, etc.). Also in the above derivations we have used a number of contradictory assumptions to make the computations tractable. For example, we assumed that the current through the wire was uniformly distributed throughout the wire in determining the inductance of the wire per unit length. However, the current for higher frequencies tends to concentrate more closely to the outside surface of the conductor due to skin effect. Also the equation for the capacitance per unit length given by (2-26) is only accurate for $D \gg r_{\omega 1}$ and $D \gg r_{\omega 2}$. This is due to the fact that closely spaced wires produce a nonuniform charge distribution; "proximity effect." A more accurate formula is given in [5] and [7]. Also we should point out that the above calculations for self and mutual inductance per unit length and capacitance per unit length were made assuming <u>static</u> conditions! Maxwell's equations for the time varying field case modify Ampere's Law as $\oint_{\text{closed path}} \vec{H} \cdot \vec{dl} = \int_{\text{closed surface}} (\vec{J} + \frac{\partial \vec{D}}{\partial t}) \cdot \vec{da}$ where the term $\vec{J}$ is the usual conduction current density and $\frac{\partial \vec{D}}{\partial t}$ is the displacement current density. For static fields, the equation reduces to Ampere's Law. For non-static conditions, i.e., sinusoidal excitation of the line especially for high frequencies, one should be very careful to determine the correct per unit length parameters and not assume that static parameters can be used to describe non-static conditions. Despite these approximations and inconsistencies many very accurate results are obtainable. Actually one could determine the per



$i_1(x,t)$
$\ell_1\Delta x$
$r_1\Delta x$
$m\Delta x \frac{\partial i_2(x,t)}{\partial t}$
$i\,(x+\Delta x,t)$
$v\,(x,t)$
$c\Delta x$
$g\Delta x$
$v(x+\Delta x,t)$
$m\Delta x \frac{\partial i_1(x,t)}{\partial t}$
$i_2(x,t)$
$\ell_2\Delta x$
$r_2\Delta x$
$i_2(x+\Delta x,t)$
Δx

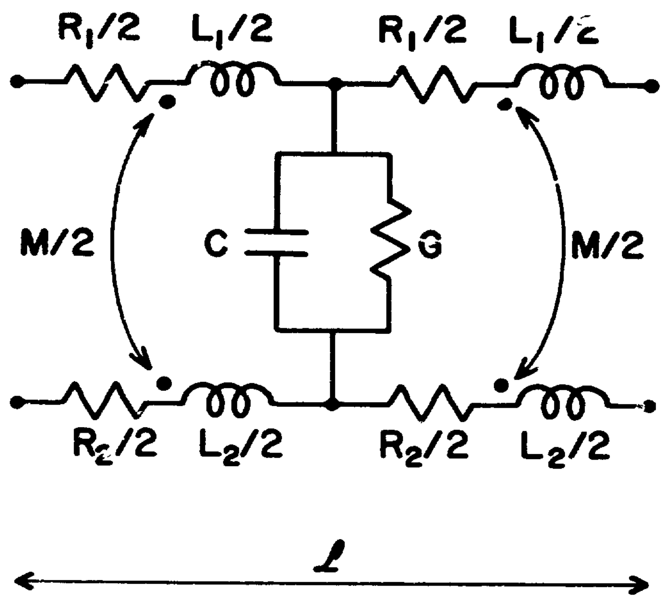
FIGURE 13

unit length parameters through measurements on representative lines using the above mathematical model without resorting to a strict field theory derivation. The derivation of these quantities for actual lines is a very difficult matter (if one does it properly).

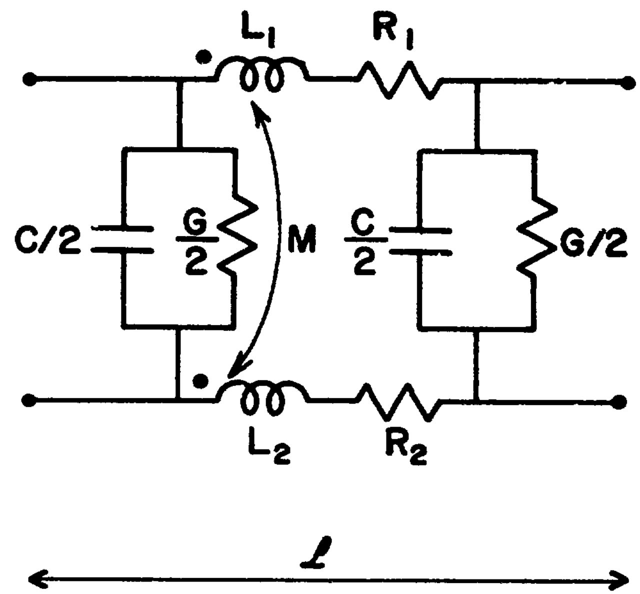
A common way of approximating the distributed parameter model of a two wire line is the following. If $\mathcal{L}$ is the total length of the line, then define $C = c \times \mathcal{L}$, $L_i = \ell_i \times \mathcal{L}$, $R_i = r_i \times \mathcal{L}$, $G = g \times \mathcal{L}$, and $M = m \times \mathcal{L}$ which yield the <u>lumped</u> equivalents given by the lumped T equivalent in Figure 14a and the lumped π (pi) equivalent in Figure 14b. One can achieve better approximations by cascading the above sections to yield the lumped double-T equivalent in Figure 15a and the lumped double-π equivalent in Figure 15b. One could continue cascading these lumped sections and in the limit as the number of sections approach infinity we would have the exact distributed model. (Notice that the <u>total</u> line parameters, L_i, R_i, G, C, M, are proportioned among the number of lumped elements in each representation.)

Let us apply the lumped equivalents to the network of Figure 3. Using the single section lumped π network in Figure 14b we have the lumped equivalent in Figure 16. In Figure 16 we have a simple lumped network to solve for the ratio $H(\omega) = V_L/V_S$. We may use standard lumped network theory to solve this problem. (We may write Node or Loop equations or we may assume $V_L = 1$ volt $\angle 0^\circ$ and work backwards to obtain V_S). Note that with this method we do not need to apply boundary conditions to determine A and B in (2-11) as in (2-12). However, this method is an approximation to the exact distributive model.

The crucial problem here is the application of these concepts to large networks of wires and their coupling affects. What must be kept in mind is the trade off between computational difficulty and accuracy of model representation. A fundamental assumption that seems to be necessary is that all wire voltages

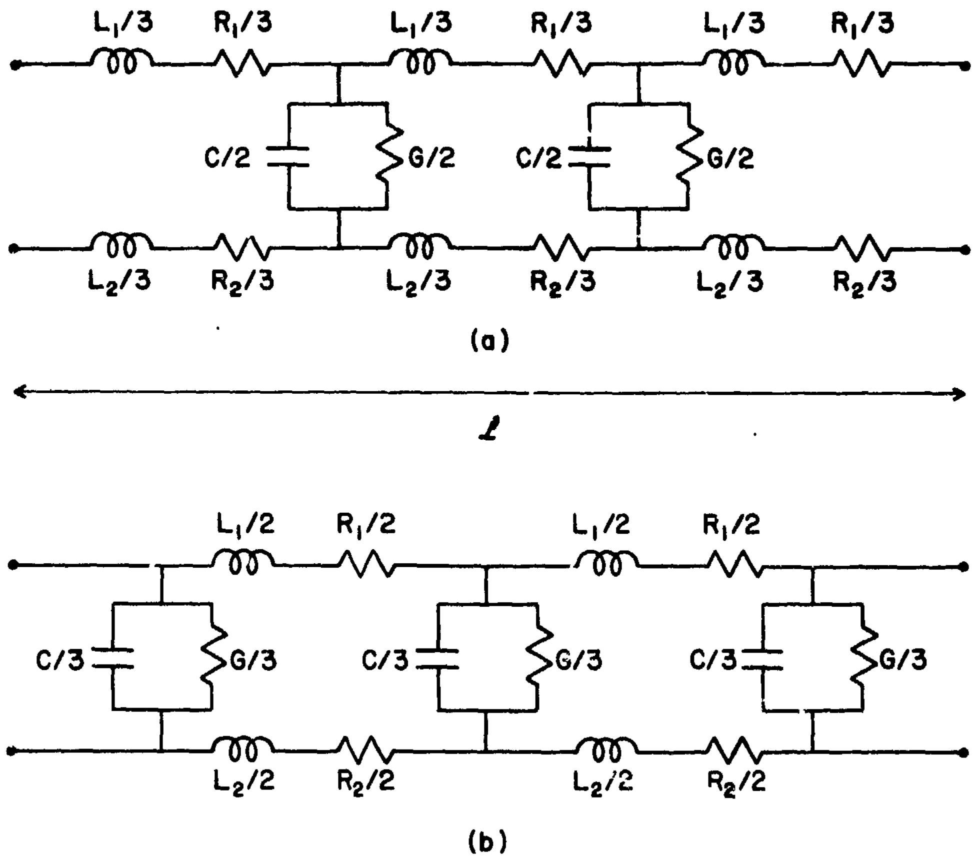


(a)



(b)

FIGURE 14.


$L_1/3$
$R_1/3$
$L_1/3$
$R_1/3$
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$L_2/2$
$R_2/2$
(b)

FIGURE 15

and currents are sinusoidal. Without going into the details of the justification we may point out that the equations for wire coupling are linear and time-invariant. Maxwell's equations are linear partial differential equations if the permittivity and permeability of the medium (ϵ and μ) do not depend on the electric field intensity, E, and the magnetic field intensity, H, respectively. Therefore a frequency domain description is equivalent to a time domain description through the Fourier Transform since superposition (for linear circuits) holds. This simplifies the required mathematics considerably. However, as shown above we still will be faced with the solution of ordinary (instead of partial) differential equations (see equation (2-5)).

A common and quite reasonable approximation to circumvent the need for differential equations is the use of lumped π or lumped T equivalent networks for representing these lines and their coupling effects. As we are aware of, lumped networks yield accurate results only for frequencies where the actual circuit dimensions are "electrically short," i.e., for dimensions much less than a wavelength. These lumped equivalents have been used for the past 50 years in the Power Industry but there the frequency(s) of interest is merely 60 HZ and the regions of applicability of the lumped models can be determined accurately. In our case, however, we will be considering frequencies up to and possibly above the GHZ range. Therefore for higher frequencies, one could use the double π, triple π, etc. to give more accurate results. Keep in mind, however, that these lumped equivalent networks are <u>approximations</u> to the <u>exact</u> distributive analysis. To provide a correlation between experimental results and mathematically predicted results one must be careful to use the appropriate model. For low frequencies, the lumped equivalents may yield good results. For high frequencies, it may be necessary to use the exact distributive model.

An exact analysis of coupling for multi-conductor lines can be made in the following straight forward manner using an exact distributive model. Consider the set of n wires shown in Figure 17 all coupled over a distance ℓ, i.e., all lines are parallel over the distance ℓ. Suppose each wire has a self inductance and resistance per unit length denoted by ℓ_i and r_i respectively. Suppose wire i has mutual inductance per unit length, m_{ij} with each of the other wires and mutual capacitance and conductance per unit length c_{ij} and g_{ij} between wire i and each of the other wires. We then may draw the equivalent circuit for an infinitesimal length Δx as shown in Figure 18. The line voltages $V_{ij}(x,t)$ and line currents $I_i(x,t)$ for $i,j = 1,\ldots,n$ $i \neq j$ are defined in Figure 17.

Let us assume that all line voltages and currents are sinusoids, i.e., $V_{ij}(x,t) = V_{ij}(x)e^{j\omega t}$ and $I_i(x,t) = I_i(x)e^{j\omega t}$. Therefore we may write

$$\begin{aligned} V_{ij}(x+\Delta x) - V_{ij}(x) &= z_{i1}\,\Delta x\, I_1(x) + \ldots + z_{ii}\Delta x\, I_i(x) + \ldots + z_{in}\,\Delta x\, I_n(x) \\ I_i(x+\Delta x) - I_i(x) &= Y_{i12}\,\Delta x\, V_{12}(x+\Delta x) + \ldots + Y_{i(n-1)n}\, V_{(n-1)n}(x-\Delta x) \end{aligned} \tag{2-49}$$

Dividing by Δx and letting $\Delta x \to 0$ we have

$$\begin{aligned} \frac{dV_{ij}(x)}{dx} &= z_{i1}\, I_1(x) + \ldots + z_{in}\, I_n(x) \\ \frac{dI_i(x)}{dx} &= Y_{i12}\, V_{12}(x) + \ldots + Y_{i(n-1)n}\, V_{(n-1)n}(x) \end{aligned} \tag{2-50}$$

where the z_{ij} and $Y_{i\alpha\beta}$ are complex impedances which are functions of ω where $\omega = 2\pi f$ and f is the frequency of interest. These quantities are obtainable in a tedious but straight forward manner from Figures 17 and 18.

Equation (2-50) leads to a set of first order differential equations in matrix notation as

$$\begin{bmatrix} \dot{\underline{V}} \\ \dot{\underline{I}} \end{bmatrix} = \begin{bmatrix} {}_{(n-1)}\underset{\sim}{0}_{(n-1)} & \underset{\sim}{z} \\ \underset{\sim}{Y} & {}_{n}\underset{\sim}{0}_{n} \end{bmatrix} \begin{bmatrix} \underline{V} \\ \underline{I} \end{bmatrix} \tag{2-51}$$

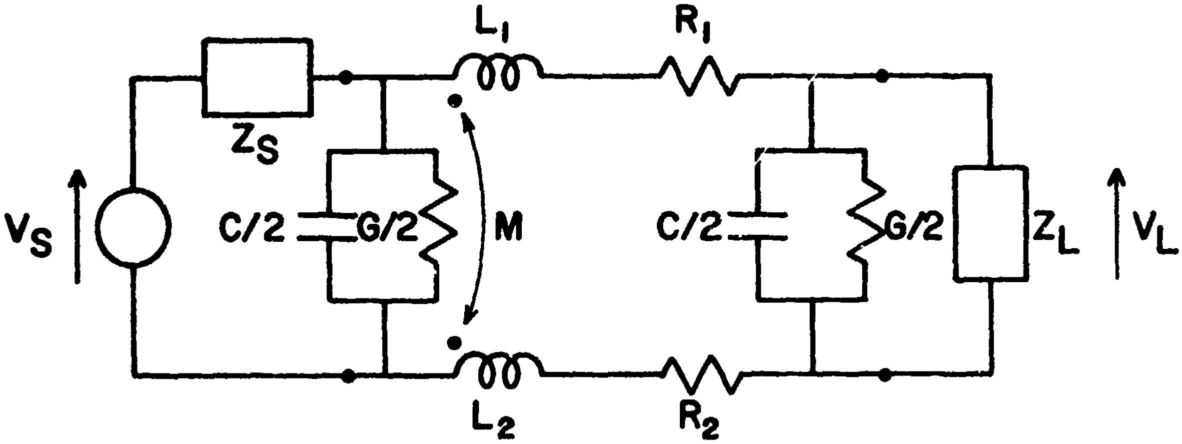


FIGURE 16

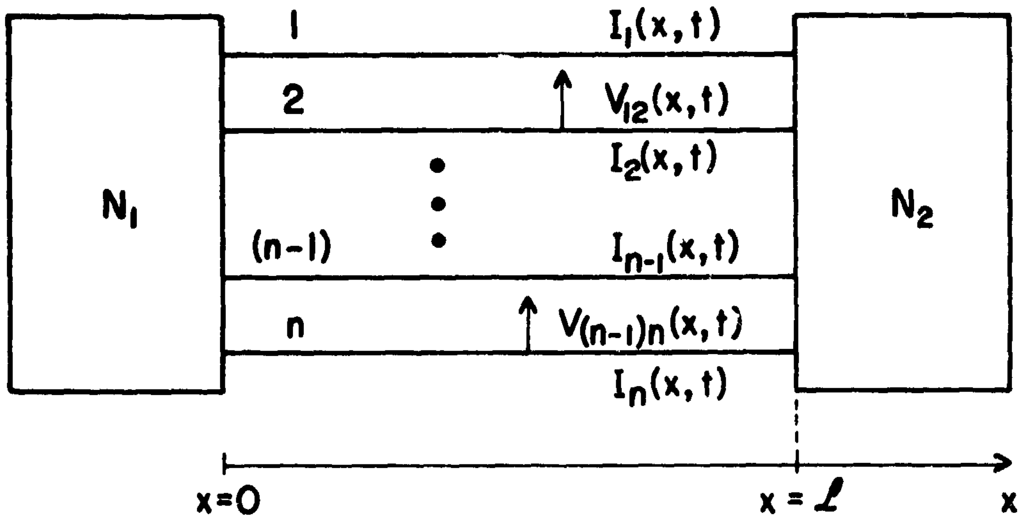


FIGURE 17

where ${}_i0_j$ is the i x j zero matrix with

$$\underline{\dot{V}} = \frac{d}{dx}\underline{V} \qquad \underline{V} = \begin{bmatrix} V_{12}(x) \\ V_{23}(x) \\ \vdots \\ V_{(n-1)n}(x) \end{bmatrix} \qquad \underline{I} = \begin{bmatrix} I_1(x) \\ I_2(x) \\ \vdots \\ I_n(x) \end{bmatrix} \tag{2-52}$$

$$\underline{\dot{I}} = \frac{d}{dx}\underline{I}$$

and the (n-1) x n matrix $\underset{\sim}{Z}$ and the n x (n-1) matrix $\underset{\sim}{Y}$ are given by

$$\underset{\sim}{Z} = \begin{bmatrix} Z_{11}\cdots\cdots\cdots Z_{1n} \\ Z_{21}\cdots\cdots\cdots Z_{2n} \\ \vdots \\ Z_{(n-1)1}\cdots\cdots Z_{(n-1)n} \end{bmatrix} \qquad \underset{\sim}{Y} = \begin{bmatrix} Y_{112}\cdots\cdots\cdots Y_{1(n-1)n} \\ \vdots \\ Y_{n12}\cdots\cdots\cdots Y_{n(n-1)n} \end{bmatrix} \tag{2-53}$$

It is well known from the theory of state-variable equations that the general solution to (2-51) is given by

$$\underline{V}(x) = \underline{V}_1 e^{\gamma_1 x} + \ldots + \underline{V}_{2n-1} e^{\gamma_{2n-1} x}$$

$$\underline{I}(x) = \underline{I}_1 e^{\gamma_1 x} + \ldots + \underline{I}_{2n-1} e^{\gamma_{2n-1} x} \tag{2-54}$$

where the (n-1) x 1 vectors $\underline{V}_i$ and the n x 1 vectors $\underline{I}_i$ for i = 1, ..., 2n-1 are arbitrary. The (2n-1) complex numbers γ_i i = 1,...,2n-1 are eigenvalues and are determined from

$$\left| \lambda_i \underset{\sim}{I}_{2n-1} - \begin{bmatrix} {}_{(n-1)}\underset{\sim}{0}_{(n-1)} & \underset{\sim}{Z} \\ \underset{\sim}{Y} & {}_n\underset{\sim}{0}_n \end{bmatrix} \right| = 0 \tag{2-55}$$

where $\underset{\sim}{I}_{2n-1}$ is the (2n-1) x (2n-1) identity matrix[17-19].

To evaluate these constant vectors in (2-54), we apply boundary conditions determined by the termination networks N_1 and N_2 to obtain $V_{12}(0)$, $V_{12}(\mathcal{L}), \ldots, V_{(n-1)n}(0)$, $V_{(n-1)n}(\mathcal{L})$, $I_1(0)$, $I_1(\mathcal{L}), \ldots, I_n(0)$, $I_n(\mathcal{L})$.

There are various computerized methods of determining the above eigenvalues and vectors. The above type of solution for multiconductor transmission lines is a natural extension of the two-wire case which we discussed previously and has been used in various forms[12-16,42-47].

If this exact (within the confidence of the per unit length line parameters) analysis is judged to be too difficult or consumes too much computational time, then an approximate lumped analysis can be used via the lumped π (pi) or lumped T networks with various numbers of stages in cascade as in Figures 14, 15 and 16. Keep in mind that this will be an approximate analysis especially for higher frequencies.

There are many other approaches to cable coupled interference which use various assumptions. In general, these approaches use lumped circuit concepts for coupling between wire pairs and give valid results only for low frequencies (or equivalently "electrically short" circuit dimensions)[20-23].

2.2 Field-to-wire and wire-to-field

Obviously, wires will be susceptible to electric and magnetic fields generated by antennas and metallic boxes. What is needed is a model for determining this vulnerability of wires and the effect of wire radiation on antennas and boxes. We have considered and modeled the effect of wire radiation (or coupling) on other wires. Since antennas and metallic boxes emit and are susceptible to electric and magnetic fields, it is sufficient to determine the fields produced by wires and the susceptibility of wires to external fields.

2.2.1 Field-to-wire

Let us first consider the field-to-wire problem. Let us suppose that we have a pair of wires (a transmission line) with an incident electric and magnetic field as shown in Figure 19. We will give two derivations of the equations

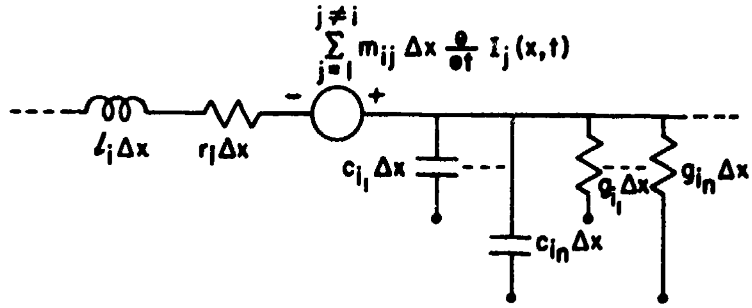


FIGURE 18

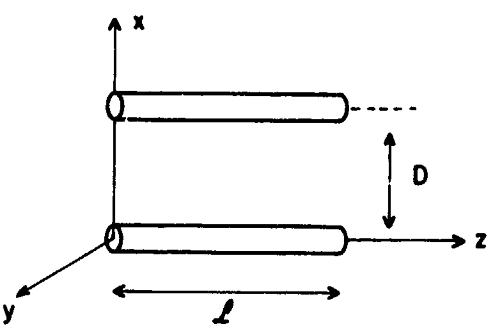


FIGURE 19

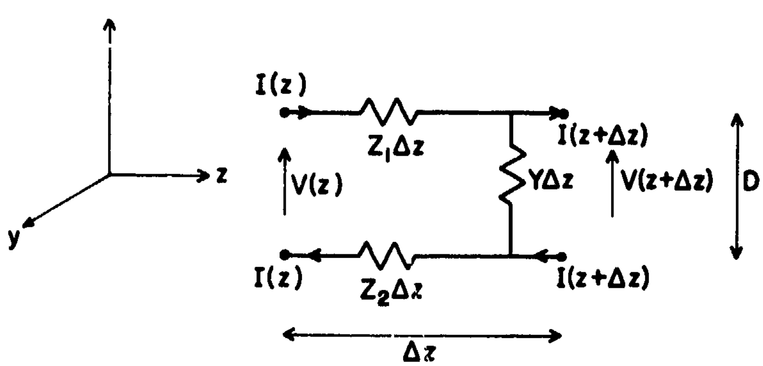


FIGURE 20

which relate the incident fields to the voltages and currents on the lines. The first is given in [24]. This derivation takes into account the distributed nature of the transmission lines and is exact in this sense. The second derivation uses simple field relations and lumped circuit concepts. Other allied results appear in [25-27]. A great deal of interest in the field-to-wire coupling problem is generated by the need for determining EMP effects on electronic circuitry.

First, consider Maxwell's equations in point form[7].

$$\begin{aligned} \nabla \times \vec{H} &= \frac{\partial \vec{D}}{\partial t} + \vec{J} \\ \nabla \times \vec{E} &= -\frac{\partial \vec{B}}{\partial t} \\ \nabla \cdot \vec{D} &= e \\ \nabla \cdot \vec{B} &= 0 \end{aligned} \tag{2-56}$$

Stokes Theorem states that

$$\int_{\text{surface}} \nabla \times \vec{A} \cdot \vec{da} = \oint \vec{A} \cdot \vec{d\ell} \tag{2-57}$$

i.e., integrating the curl of a vector quantity over a surface is equivalent to obtaining the line integral of the vector quantity along the closed perimeter enclosing the surface. The direction of the line integral is such that the direction of da complies with the right hand rule. Integrating (2-56) in accordance with (2-57) produces Maxwell's equations in integral form as

$$\begin{aligned} \oint \vec{H} \cdot \vec{d\ell} &= \int_S \left(\frac{\partial \vec{D}}{\partial t} + \vec{J} \right) \cdot \vec{da} \\ \oint \vec{E} \cdot \vec{d\ell} &= -\int_S \frac{\partial \vec{B}}{\partial t} \cdot \vec{da} \end{aligned} \tag{2-58}$$

$$\int_{\substack{\text{Closed}\\\text{Surface}\\\text{Enclosing}\\\text{Volume}}} \vec{D} \cdot \vec{da} = \int_{\text{Volume}} e^{dv}$$

$$\int_{\substack{\text{Closed}\\\text{Surface}\\\text{Enclosing}\\\text{Volume}}} \vec{B} \cdot \vec{da} = 0 \tag{2-58}$$

where the last two equations of (2-58) follow from (2-56) and Gauss's Divergence Theorem, i.e.,

$$\int_{\text{Volume}} \nabla \cdot \vec{A}\, dv = \int_{\substack{\text{Closed}\\\text{Surface}\\\text{Enclosing}\\\text{Volume}}} \vec{A} \cdot \vec{da} \tag{2-59}$$

Consider a differential length of line Δz as shown in Figure 20 where the per unit length parameters of the line are included. Assume without loss of generality the incident fields are sinusoidal and consist of components

$$\begin{aligned} E_x(x,z,t) &= E_x(x,z)e^{j\omega t} \\ E_z(x,z,t) &= E_x(x,z)e^{j\omega t} \\ By(x,z,t) &= By(x,z)e^{j\omega t} \end{aligned} \tag{2-60}$$

Integrating the second equation of (2-58) yields

$$\begin{aligned} &\int_0^D [E_x(x,z+\Delta z) - E_x(x,z)]\, dx \\ &\quad - \int_z^{z+\Delta z} [E_z(D,z) - E_z(0,z)]\, dz \\ &\quad = -j\omega \int_z^{z+\Delta z} \int_0^D By(x,z)\, dx\, dz \end{aligned} \tag{2-61}$$

We may easily identify

$$\lim_{\Delta z \to 0} \frac{V(z+\Delta z) - V(z)}{\Delta z} = \frac{dV(z)}{dz} = \lim_{\Delta z \to 0} \frac{1}{\Delta z} \int_0^D [E_x(x,z+\Delta z) - E_x(x,z)]\, dx$$

$$\lim_{\Delta z \to 0} \frac{1}{\Delta z} \int_z^{z+\Delta z} [E_z(D,z) - E_z(0,z)]\, dz = ZI(z) \tag{2-62}$$

where $Z = Z_1 + Z_2$ and therefore have the relation

$$\frac{dV(z)}{dz} + ZI(z) = j\omega \int_0^D B_y(x,z)\, dx \tag{2-63}$$

Similarly we may obtain

$$\frac{dI(z)}{dz} + YV(z) = - Y \int_0^D E_x(x,z)\, dx \tag{2-64}$$

From these equations, one may obtain

$$\frac{d^2V(z)}{dz^2} + ZYV(z) = ZY \int_0^D E_x(x,z)\, dx \tag{2-65}$$

and similarly

$$\frac{d^2I(z)}{dz^2} - ZYI(z) = - Y[E_z(D,z) - E_z(0,z)] \tag{2-66}$$

Equations (2-65) and (2-66) may be solved in standard fashion after integrating the source field $E_x(x,z)$ and each solution will contain two undetermined constants in the homogeneous part of the solution. These constants may be evaluated with the boundary conditions $V(0)$, $V(\ell)$, $I(0)$ and $I(\ell)$ once one specifies the terminations for the line. The result will yield $V(z,t) = V(z)e^{j\omega t}$ and $I(z,t) = I(z)e^{j\omega t}$.

A rather crude approximation can be obtained in the following way. This solution neglects the scattering effect of the transmission line. We know that a changing magnetic field will induce potential difference in a loop which is proportional to the rate of change of flux enclosed by the loop, i.e.,

$$V_m = - \frac{d\psi}{dt} = -j\omega \int_0^{\ell} \int_0^D B_y(x,z)\, dx\, dz \tag{2-67}$$

Let us use the lumped π equivalent circuit for the transmission line and include V_m in the loop as shown in Figure 21. In Figure 21, the per unit length parameters

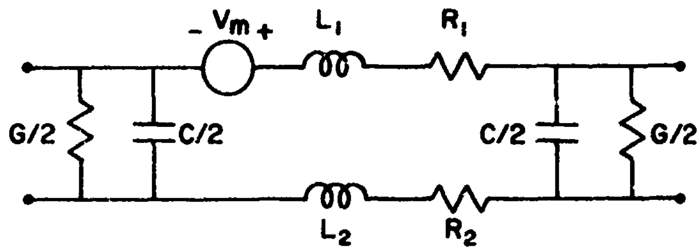
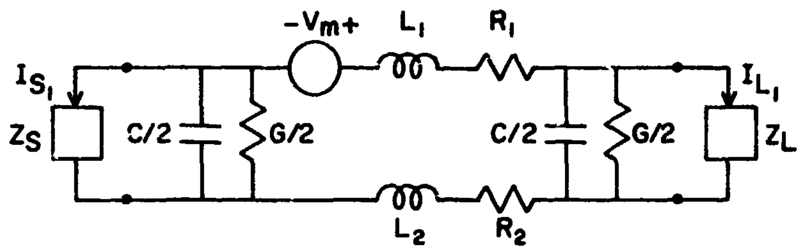
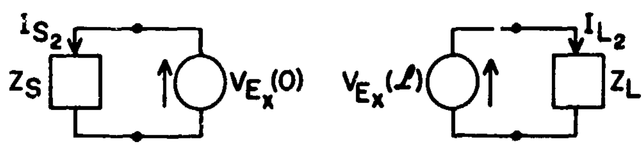


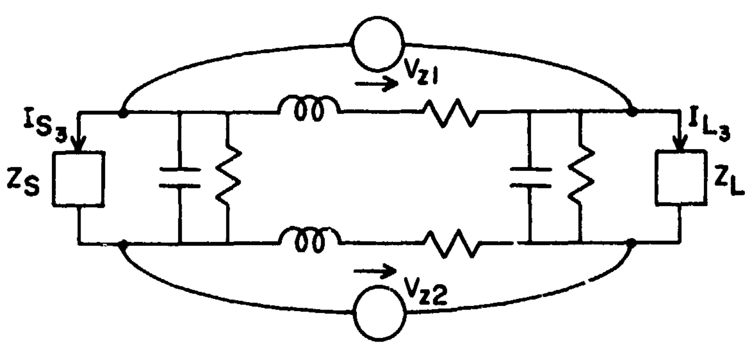
FIGURE 21



(a)



(b)



(c)

FIGURE 22

of the lines are ℓ_1, ℓ_2, r_1, r_2, c,g and $L_i = \ell_i \times \mathcal{L}$, $R_i = r_i \times \mathcal{L}$, $C = c \times \mathcal{L}$, $G = g \times \mathcal{L}$.

Now let us consider the effects of E_x and E_z separately. We may write

$$
\begin{aligned}
V_{E_x}(0) &= -\int_0^D E_x(x,0)\,dx \\
V_{E_x}(\mathcal{L}) &= -\int_0^D E_x(x,\mathcal{L})\,dx
\end{aligned}
\qquad (2\text{-}68)
$$

The electric field $E_z(x,z)$ will produce voltages

$$
\begin{aligned}
V_{z_1} &= -\int_0^{\mathcal{L}} E_z(D,z)\,dz \\
V_{z_2} &= -\int_0^{\mathcal{L}} E_z(0,z)\,dz
\end{aligned}
\qquad (2\text{-}69)
$$

Now suppose (for example) the termination impedances are Z_S and Z_L. Since the equations of the line and Maxwell's equations are linear (for linear media) then we may superimpose the effects of By, E_x, and E_z to find the currents in each of the termination impedances as

$$
\begin{aligned}
I_L &= I_{L_1} + I_{L_2} + I_{L_3} \\
I_S &= I_{S_1} + I_{S_2} + I_{S_3}
\end{aligned}
\qquad (2\text{-}70)
$$

which are defined in Figure 22. The accuracy of the above approximate solution has not been determined through experiment. For "electrically short" lines, however, this approximation may yield reasonably accurate results with an associated computational advantage over the exact analysis.

2.2.2 Wire-to-field

Here we will only discuss the concepts. It is well known that a current carrying wire will produce a magnetic field. This magnetic field can be related to the current through the Biot-Savart Law[6-8]. Also the potential difference between two wires of a transmission line or between two points along a transmission

line will give rise to a radiated electric field. Various approximations may be used to model the transmission line as a wire-type antenna to obtain the resulting radiated fields. The exact solutions for radiating wires are well known once one is given the current distribution along the wire. This piece of information is usually the most difficult to obtain in the solution of antenna problems.

Suppose we are given the source and load terminations for a two-wire transmission line as in Figure 3. An exact solution for the current on each wire as a function of distance along the wire can be found in a straightforward manner as given by (2-11). Consequently, we may solve any multiconductor transmission line problem for the currents along each wire and use relatively straightforward integral field equations to derive the spatial electric and magnetic fields. We then may superimpose the fields due to each wire separately at each point in space to obtain the electromagnetic field profile due to these radiating wires. In theory then, this problem is quite straightforward in solution. In practice, one would probably approximate the solution at desired points in space.

2.3 Antenna radiation and reception

If we determine the electromagnetic fields produced by an antenna then we may determine the effect of this source upon wires through the analysis in 2.2. At this point, we may invoke reciprocity to determine the reception capabilities of an antenna for incident fields. To simplify matters, we will discuss antenna-to-antenna coupling through some methods which have been used a great deal in the past.

The exact solution for field-to-antenna, antenna-to-field and antenna-to-antenna coupling has only been solved in general, closed form for filamentary wire-type antennas (loop, dipole and long wire). Consequently, more general

antennas yield solutions more easily on an "input-output" basis.

For determining the coupling between two antennas, we could assume as a highly simplified but often used approximation that both antennas are isotropic point source radiators as shown in Figure 23. With this assumption, the radiated power is given by

$$P_T/4\pi R^2 = \vec{P} \; (\text{watts}/\text{m}^2) \tag{2-71}$$

where $\vec{P}$ is the power per unit area passing through a sphere of area $4\pi R^2$ and the transmitter is producing P_T watts of power. In this case the Poynting vector becomes

$$\vec{P} = \vec{E} \times \vec{H} \tag{2-72}$$

and since we are assuming plane waves in free space (i.e., isotropic point source)

$$|\vec{E}| = z \; |\vec{H}| \tag{2-73}$$

where $|\vec{E}|$ is the magnitude of $\vec{E}$ and z is the impedance of the transmission medium (377 ohms for free space). Thus we have

$$P_T/4\pi D^2 = \frac{|\vec{E}|^2}{z} \tag{2-74}$$

or

$$|\vec{E}| = \sqrt{\frac{zP_T}{4\pi D^2}} \tag{2-75}$$

as the magnitude of the electric field in the vicinity of the receiving point source antenna. We could also convert this electric field to an equivalent voltage, V, at the base of the receiving antenna using another highly simplified (but often used) approximation with an "effective antenna height" $\hat{h}$ as

$$V = \hat{h} \; |\vec{E}| \tag{2-76}$$

Clearly this is a very unsatisfactory model for reasons almost too obvious to mention. The model does not consider the peculiarities of the particular antenna, such as beam pattern, etc. Also the "effective height" of the receiving

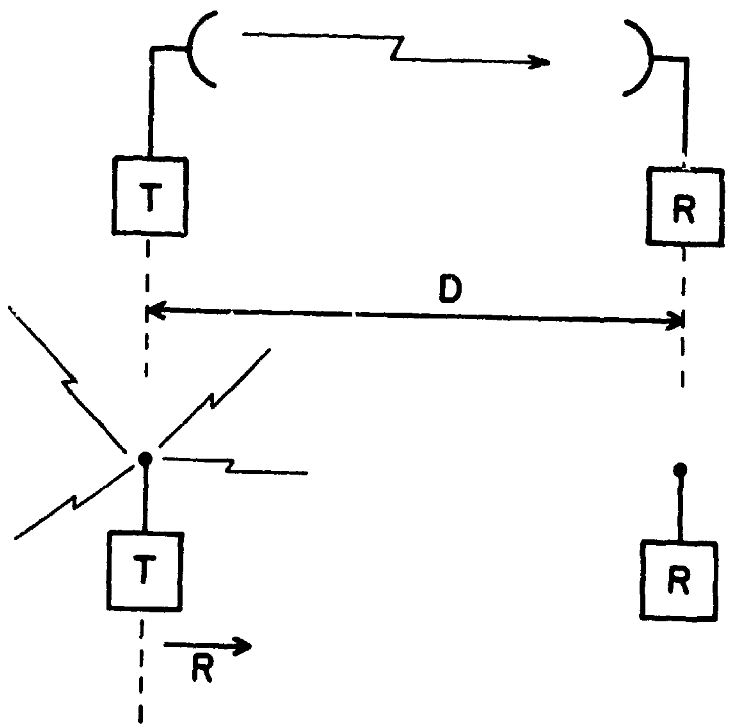


FIGURE 23

antenna, $\hat{h}$, can be directly calculated only for antennas of very simple geometries. In most cases this quantity is very subjective and is obtainable, at best, through direct measurement. The model, however, is the simplest quantitative model of antenna coupling.

Define the effective area of an antenna as the ratio of the received power of an antenna to the power density of the incoming wave, i.e.,

$$A = \frac{P_R}{P_{in}} \tag{2-77}$$

Suppose we have a transmitting antenna of effective area A_T and a receiving antenna of effective area A_R. A commonly used formula for determining the received power of an antenna is the Friis transmission formula[28, 29, 30]

$$P_R = \frac{P_T\, A_T\, A_R}{D^2 \lambda^2} \tag{2-78}$$

where $\lambda = {}^{2\pi}/f$ and f is the frequency of transmission. It should be noted that in the above usage, the effective areas A_T and A_R must be given in the direction of transmission. This formula has been used extensively today even though it was determined in 1946. The reason for its success seems to be its simplicity and the very general definition in (2-69). The burden is placed upon the user to determine precisely what the effective area of each antenna is, no matter how complicated , geometrically, his antenna is.

In the case of coupling between antennas collocated on an aircraft, the trend has been to use the Friis transmission formula and a correction factor to take into account the bending of the waves as they follow the curvature of the fuselage[28, 29, 30]. Wing shading and diffraction effects are included in [28]. For propagation around noncylindrical and especially unsymmetrical bodies, the analysis could be amenable only to experimental determination.

2.4 Box radiation and reception

Very little has been done in modeling field-to-box, box-to-field and box-to-box coupling, and it is doubtful that any significant results will be obtained in the near future due to the variability in geometry and location of the boxes (which are not designed for functional coupling purposes as with antennas) within a system. This area of coupling determination seems to be the most difficult to model. This is due to at least two reasons. The shapes of boxes containing electronic equipment is not designed for functional purposes and, consequently, meager standardized information is available. Furthermore, what is required essentially is, first, the determination of the coupling to and from the box wall and external fields and, more importantly, the coupling between the box and its internal electronic circuitry (which is the fundamental area of concern).

Boxes containing 400 Hz transformers have been replaced by an equivalent magnetic dipole (see the discussion of the McDonnell Program). Other allied results are found in [33].

III. THE BOEING PROGRAM[34]

This program was procured by the Air Force Avionics Laboratory (AFAL) to analyze the intrasystem EMC of avionics systems and is concerned with analyzing cable-to-cable and direct cable conducted coupling. To illustrate the program we will use a simple example. Consider the electrical network in Figure 24.

In Figure 24, the source S represents a source of electromagnetic energy (either desired or undesired) and the receptor R represents a device which may be critically affected by electromagnetic energy (either desired, undesired or both). Let us assume that these devices are connected by parallel wire lines with ground returns separated by distance D over a common length $\mathcal{L}$.

The assumption is made that the source S, load Z_1, load Z_2 and receptor R are linear devices with respect to their input terminals. This assumption is implicit in the representation of each of these devices with a Norton (or Thevenin) equivalent circuit as shown in Figure 25.

In Figure 25, $Z_S(f)$, $Z_1(f)$, $Z_2(f)$ and $Z_R(f)$ are frequency dependent impedances, e.g., $Z(f) = R(f) + j\,X(f)$ where R(f) and X(f) are the real and imaginary parts of Z(f) respectively and are functions of frequency f. The Norton equivalent current source, I(f), for S is a frequency dependent amplitude function which is equivalent to the magnitude of the Fourier Transform of the short circuit current associated with S. The phase angle of the Fourier spectrum associated with this source is neglected. There may be serious questions in some cases about the assumption of linearity of the devices (which is inherent in the use of a Norton or Thevenin representation). However, the magnitude of the problem seemingly dictates this assumption as a necessity.

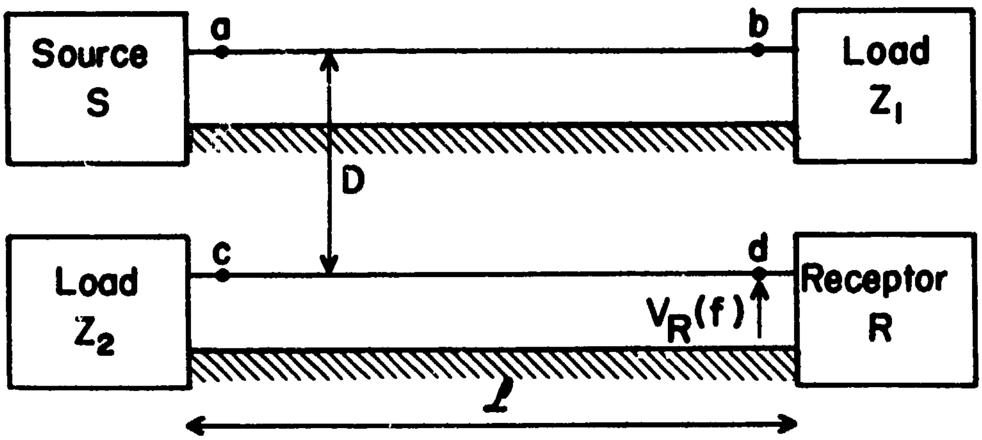


FIGURE 24

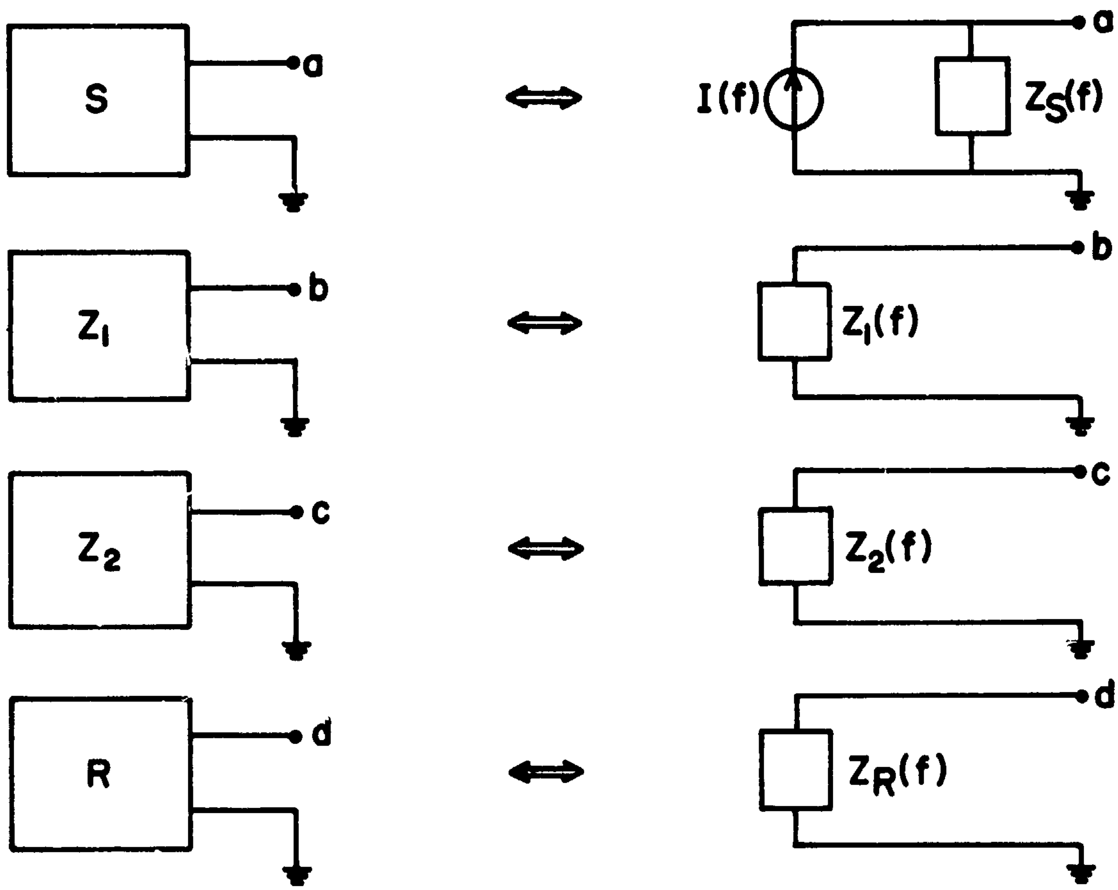


FIGURE 25

The above impedances and the spectrum of the current source are determined through direct measurements on the actual devices. The magnitudes of $Z_1(f)$, $Z_2(f)$ and $Z_R(f)$ given by $|Z_1(f)|$, $|Z_2(f)|$ and $|Z_R(f)|$ respectively are determined and an equivalent two terminal RLC type network is synthesized whose input impedance are $\hat{Z}_1(f)$, $\hat{Z}_2(f)$ and $\hat{Z}_R(f)$. The synthesized circuits are such that $|Z_1(f)| \cong |\hat{Z}_1(f)|$, $|Z_2(f)| \cong |\hat{Z}_2(f)|$, $|Z_R(f)| \cong |\hat{Z}_R(f)|$ where $\cong$ denotes "approximately equal to." The reason for determining an equivalent RLC network instead of using the plots of input impedance magnitude vs. frequency is that the data inputs to the program are simplified since only the values of the elements and the circuit configuration of the RLC equivalent need be specified in terms of an easily obtained input impedance equation. However, one could approximate the curve $|Z(f)|$ with straight line segments and determine the equation for $|\hat{Z}(f)|$ for this approximation to any desired accuracy. The equivalent current source of the source, I(f), is simply a magnitude versus frequency spectrum. The susceptibility level of the receptor is a magnitude versus frequency spectrum of the input voltage to the receptor, $V_R(f)$, and is determined external to the analysis.

The transmissions lines and their coupling effects are modeled using the single section lumped π equivalent as in Figure 26. In Figure 26, $L = \ell \times \mathcal{L}$, $C = c \times \mathcal{L}$, $C_m = c_m \times \mathcal{L}$, $M = \ell_m \times \mathcal{L}$, $R(f) = r(f) \times \mathcal{L}$ is a frequency dependent resistance function which includes skin effect and $\mathcal{L}$ is the common length of the generator and receptor transmission line.

What is required is the computation of the magnitude of $V_R(f)$ to be compared with the susceptibility spectrum of the receptor to determine if an incompatibility exists. If the magnitude of the received voltage spectrum, $V_R(f)$, exceeds the receptor susceptibility spectrum at any frequency then an incompatibility exists. The truth of the converse of this statement is of

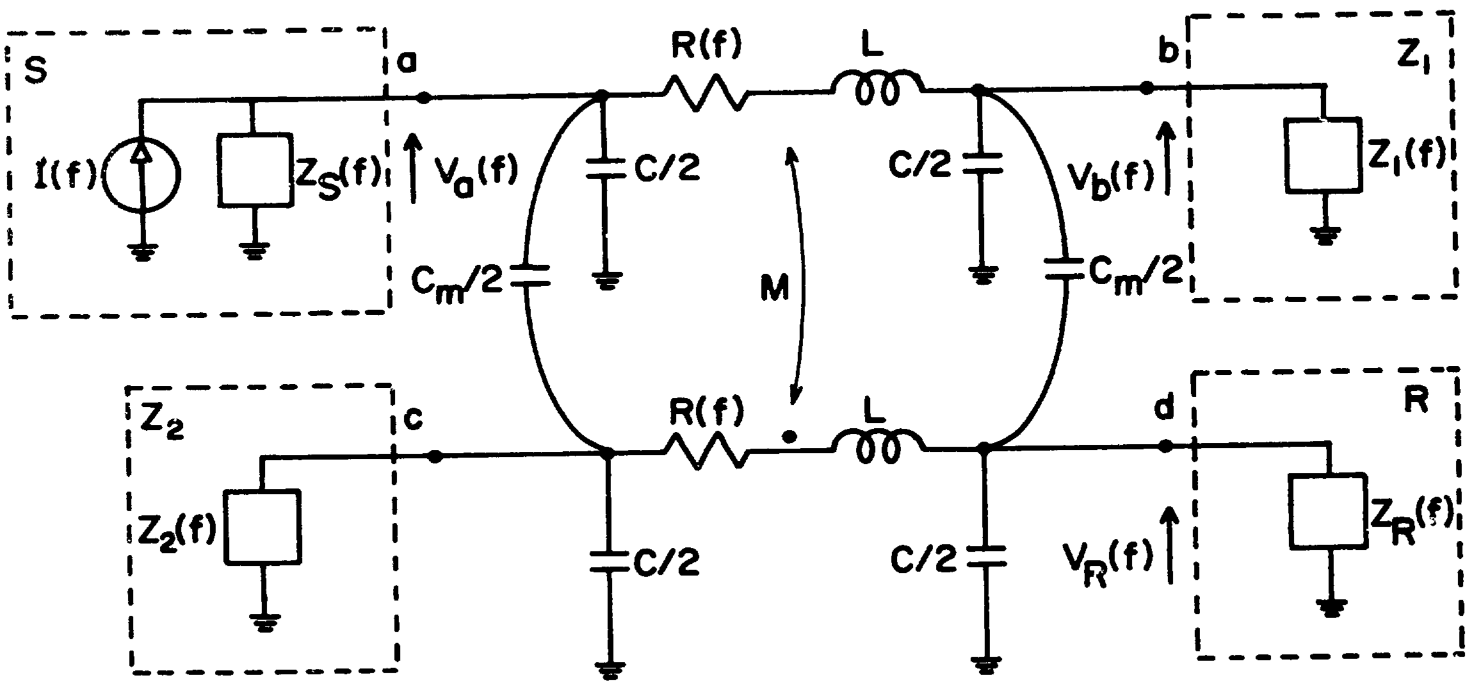


FIGURE 26

course completely dependent upon having a relevant and meaningful susceptibility spectrum.

$V_R(f)$ is computed using the assumption that the loading effect of the receptor circuit on the generator circuit is negligible. This assumption manifests itself in the following way. The voltages $V_a(f)$ and $V_b(f)$ and the current $I_1(f)$ in Figure 26 are computed from the circuit of Figure 27. With this information, the receptor voltage, $V_R(f)$, is computed from the circuit of Figure 28 where $V_m(f) = j\omega MI_1(f)$ and $\omega = 2\pi f$.

The program is applied to the computation of interference for networks with more than one set of emitter wires and/or more than one set of receptor wires in the following manner. Let us suppose that the entire cable coupling network consists of p sets of generator (or emitter) wires $G_1, G_2, \ldots, G_p$ and q sets of receptor wires $R_1, R_2, \ldots, R_q$ as illustrated in Figure 29.

The system we are considering is then represented as in Figure 30. Each generator-receptor pair is considered separately and modeled as in Figure 26. (The extension to the case without the ground plane is also considered.) A frequency dependent transfer function for each generator-receptor pair is determined as described previously (see Figure 26, 27 and 28) given by $H_{ij}(\omega)$ where the subscript i represents the generator and the subscript j represents the receptor. The reader should note that this derivation implicitly assumes <u>no</u> coupling between each generator, <u>no</u> coupling between each receptor and <u>no</u> coupling between any generator-receptor pair other than G_i and R_j (the pair being considered). Furthermore, it is assumed that receptor R_j does not "load" generator G_i in calculating the transfer function $H_{ij}(\omega)$. In other words, we are deleting all generators and all receptors except G_i and R_j in determining the transfer function $H_{ij}(\omega)$. In determining $H_{ij}(\omega)$ in this reduced network we further assume that the presence of R_j does not affect the values of the voltages

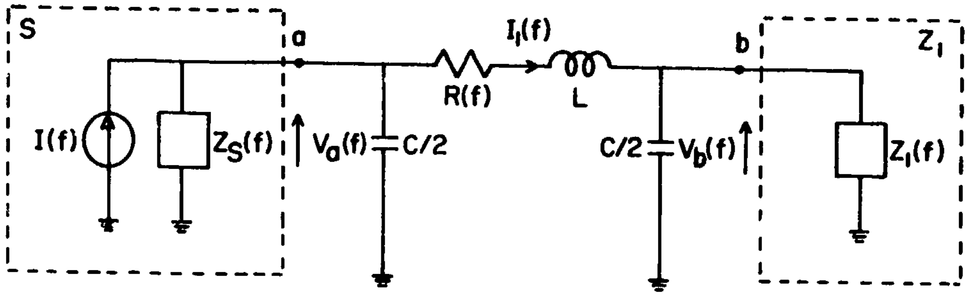


FIGURE 27

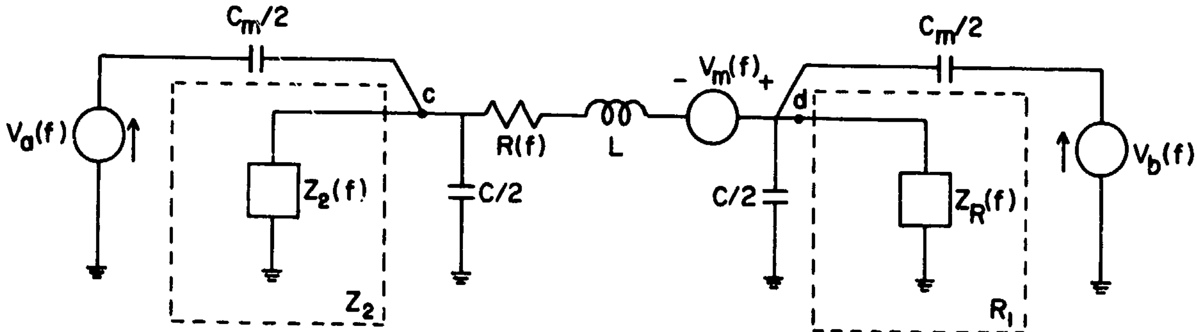


FIGURE 28

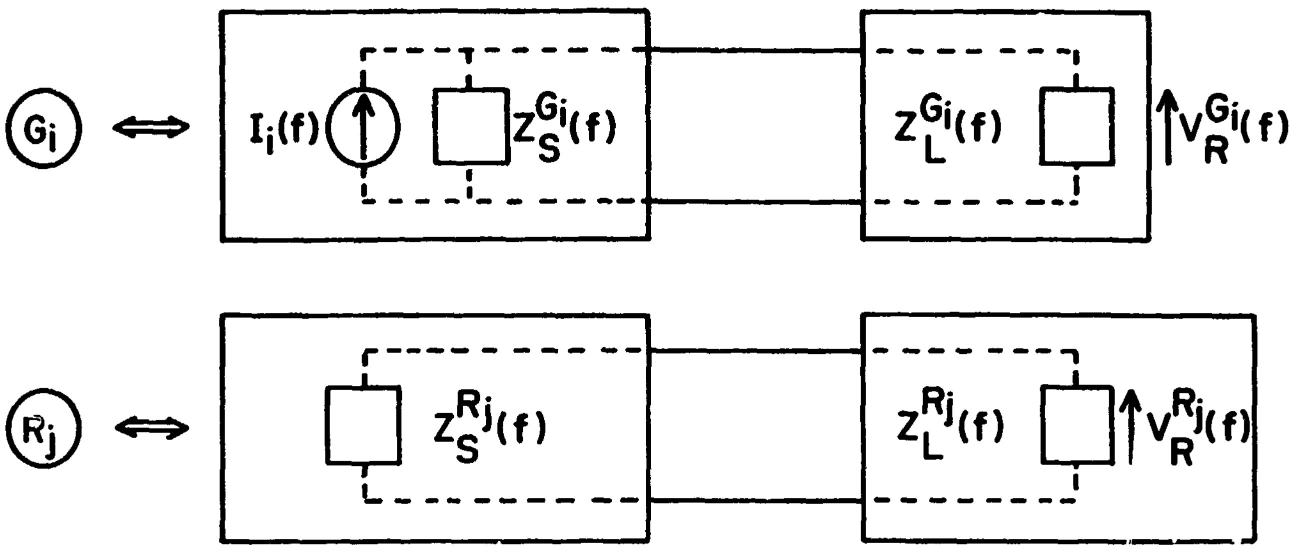


FIGURE 29

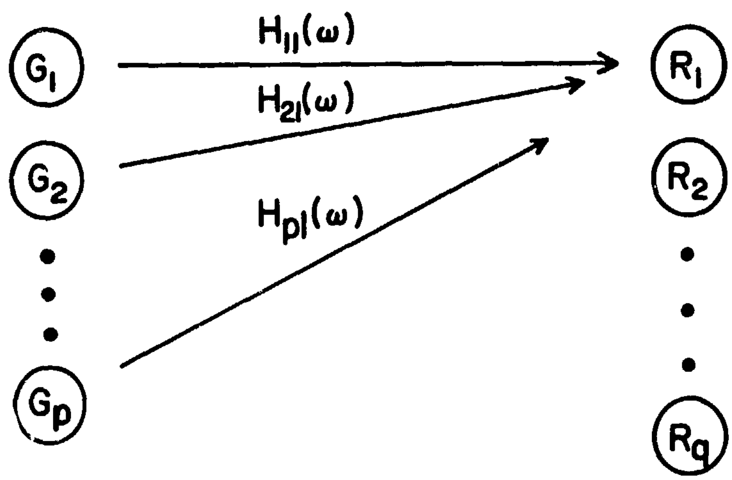


FIGURE 30

and currents in the generator network G_i. The transfer function $H_{ij}(\omega)$ then provides

$$H_{ij}(\omega) = \frac{V_R^{R_j}(\omega)}{I_i(\omega)} \qquad (3\text{-}1)$$

under the above assumptions. These approximations were apparently made to reduce computation time. To provide an exact solution (within the confidence in the line parameters and the use of the lumped π equivalent), one would have to solve a horrendous network problem by including all generator and receptor circuits with all associated cross-coupling effects. Notice that the voltage across the generator load $Z_L^{G_i}(f)$, is calculated automatically once one solves the generator network. Therefore, the direct coupled interference transfer function is given by

$$H_{ii}(\omega) = \frac{V_R^{G_i}(\omega)}{I_j(\omega)} \qquad (3\text{-}2)$$

At this point, let us discuss certain important concepts to be considered in using this program as well as any other compatibility prediction program. Obviously, in the actual system deployment, the voltage across a receptor impedance, $Z_R^{R_j}(\omega)$, will be the sum of the effects due to all generators of interference (not only wire type generators but also antenna, box and intersystem sources). One cannot simply excerise this program by determining the voltage across a wire type receptor load impedance due to each wire type source and comparing each of these voltage spectra to the susceptibility threshold spectrum of the receptor to determine compatibility and conclude that if none of the received voltage spectrums exceed the susceptibility threshold spectrum the system will be a compatible one. This useage would represent a gross conceptual error and in many cases would yield an incorrect conclusion. The receptor in an actual system will be affected to some degree by all sources (wire, box, antenna and intersystem sources.

As an example of the proper method of analysis, let us assume that the entire system consists of two wire-type sources and one transmitter with associated antenna. Also let us assume that there is only one receptor and this is a wire-type receptor. The "system" can be modelled as in Figure 31. In Figure 31, the coupling between the wire circuits represents mutual inductance and capacitance coupling and the coupling between the antenna and wire receptor is only "one-way." In this model we are assuming only that the output of the transmitter is unaffected by the rest of the system and the wire generators, G_1 and G_2, are not affected by the transmitter. Also we are assuming that there is no box radiation or reception. Notice that we are including the effects of intersystem sources. These intersystem sources will affect the entire system and should be included unless it is assumed that their effects are negligible. We could then solve the reduced systems in Figure 32, 33, 34 and 35.

In Figure 32, we simply "turn-off" the transmitter and both current sources $I_1(\omega)$ and $I_2(\omega)$ and determine the contribution to $V_{R_1}(\omega)$ due to intersystem effects only, $V_{R_1}(\omega)_1$. In Figure 33, we "turn-on" the transmitter, "turn-off" both current sources $I_1(\omega)$ and $I_2(\omega)$ and intersystem effects (in making the calculations) to determine the contribution to $V_{R_1}(\omega)$ due to the transmitter only, $V_{R_1}(\omega)_2$. Similarly, in Figures 34 and 35 we determine the contributions to $V_{R_1}(\omega)$ due to the current sources $I_1(\omega)$ and $I_2(\omega)$ only, respectively.

Notice that if a wire generator is "turned-off" we do not delete the wires from consideration. In other words, the "deactivated" wire generators still exist but do not have excitations. For example, in Figure 33, the receptor voltage $V_{R_1}(\omega)_2$ due to the transmitter will be affected not only by the transmitter output but also by the parasitic nature of the deactivated generator lines.

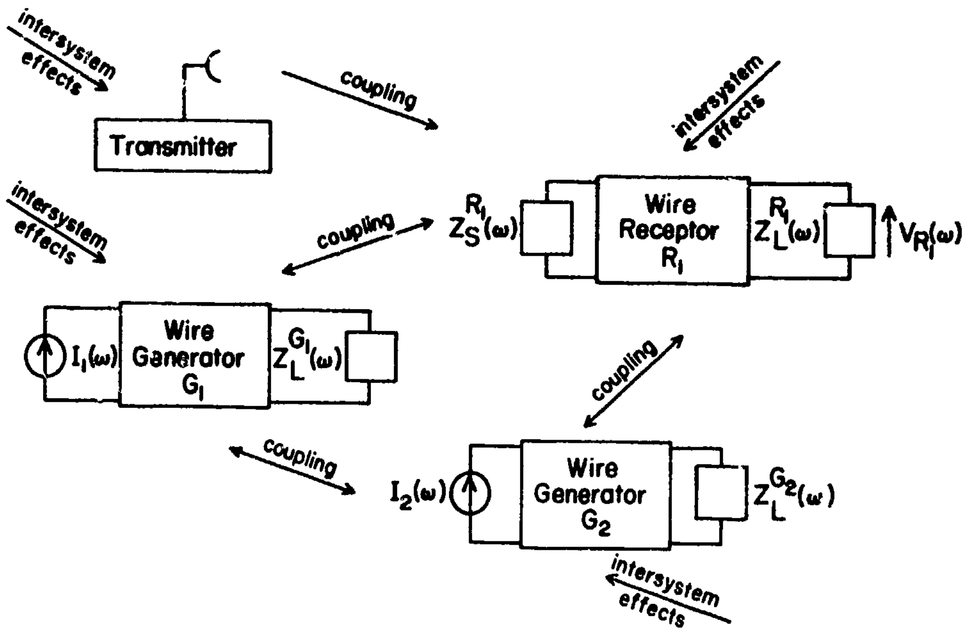


FIGURE 31

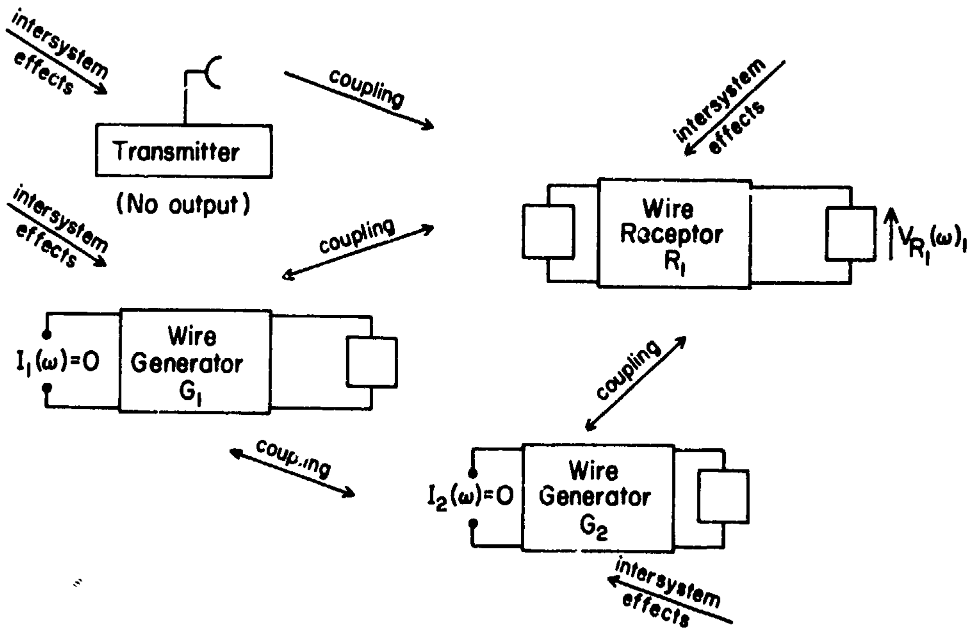


FIGURE 32

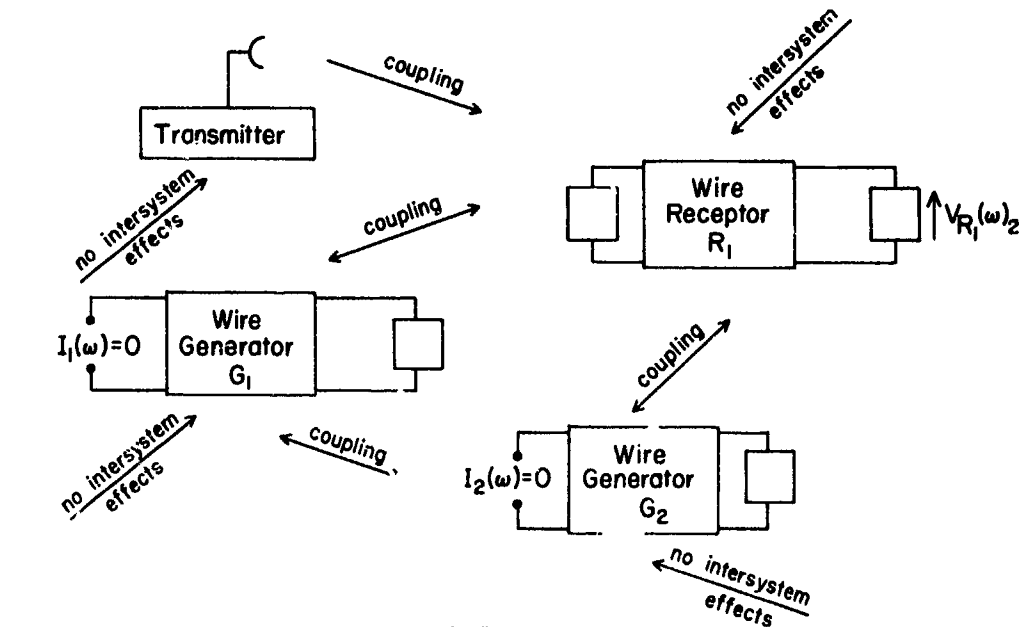


FIGURE 33

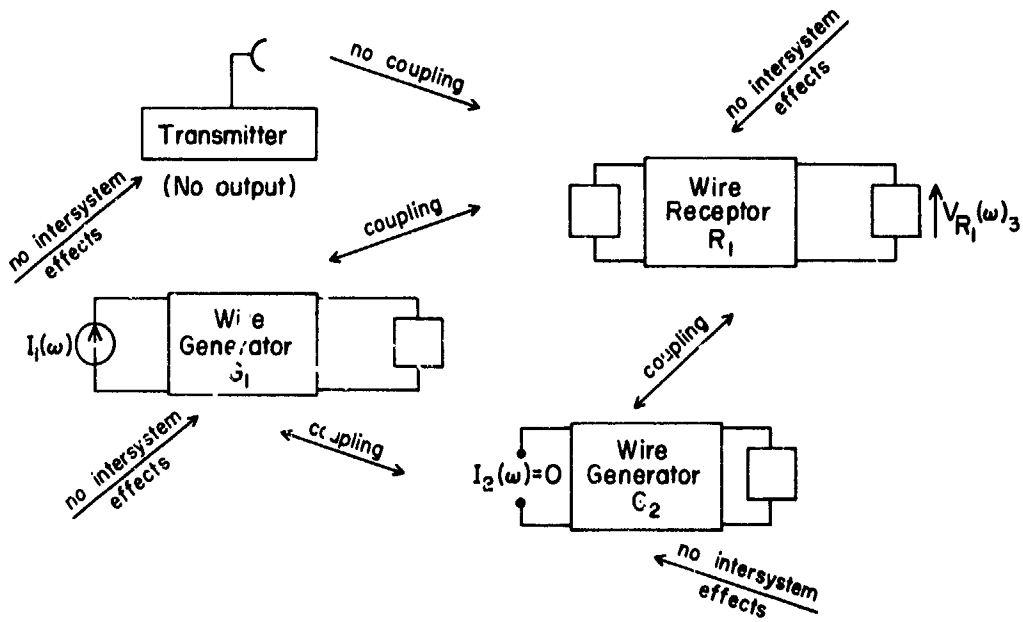


FIGURE 34

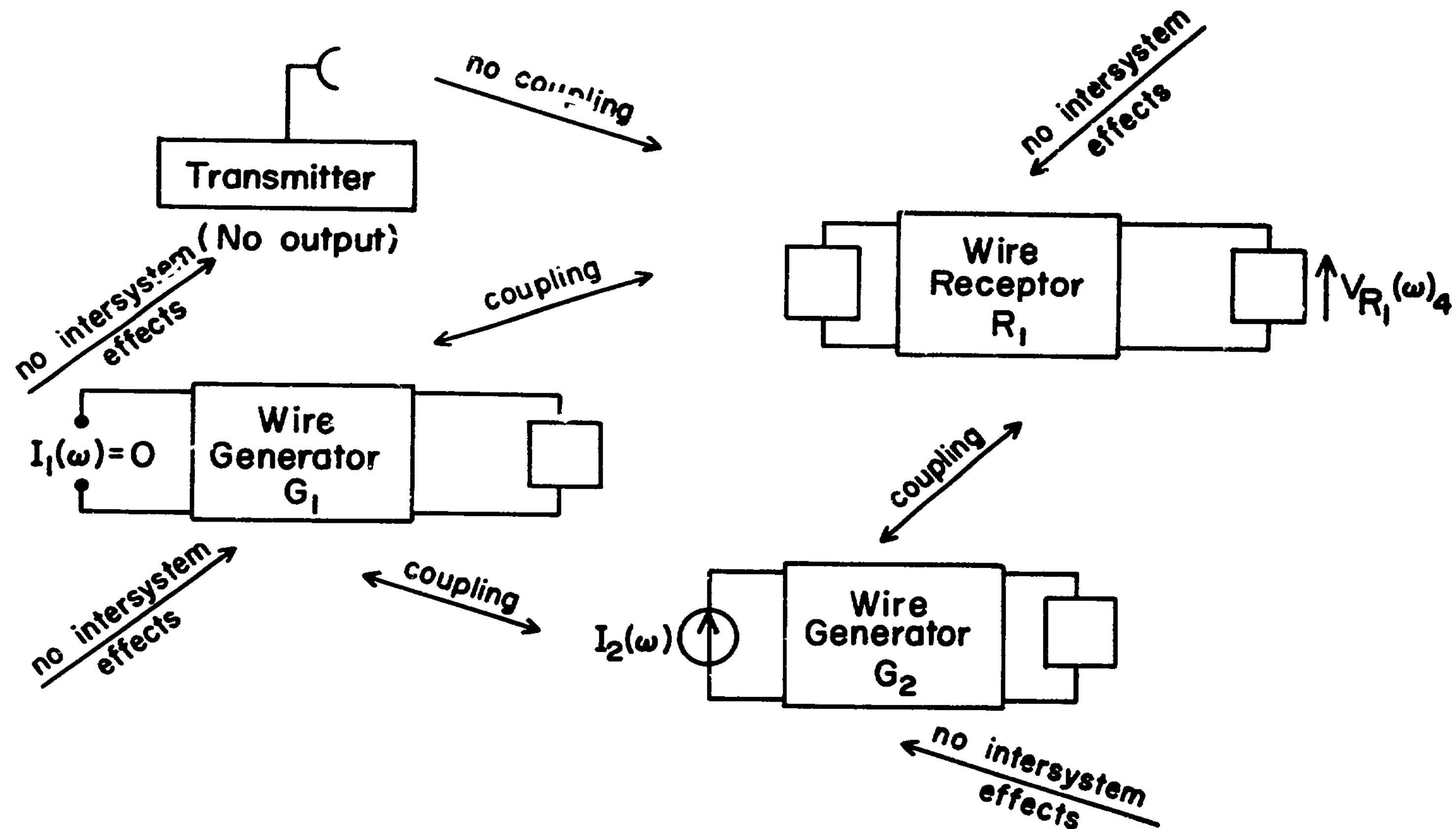

Transmitter
(No output)
no coupling
no intersystem effects
coupling
Wire Receptor R_1
$V_{R_1}(\omega)_4$
no intersystem effects
$I_1(\omega)=0$
Wire Generator G_1
coupling
no intersystem effects
coupling
$I_2(\omega)$
Wire Generator G_2
no intersystem effects

FIGURE 35

Since superposition of source effects is valid for linear coupling media (the lines are linear and the medium is linear), the voltage across the receptor load, $V_{R_1}(\omega)$, will be given by

$$V_{R_1}(\omega) = \underbrace{V_{R_1}(\omega)_1}_{\substack{\text{due to}\\ \text{inter-}\\ \text{system}\\ \text{effects}\\ \text{only}}} + \underbrace{V_{R_1}(\omega)_2}_{\substack{\text{due to}\\ \text{the}\\ \text{trans-}\\ \text{mitter}\\ \text{only}}} + \underbrace{V_{R_1}(\omega)_3}_{\substack{\text{due to}\\ G_1\\ \text{only}}} + \underbrace{V_{R_1}(\omega)_4}_{\substack{\text{due to}\\ G_2\\ \text{only}}} \tag{3-3}$$

Now if one compares the received threshold, $V_{R_1}(\omega)$, to the susceptibility threshold of R_1, a valid conclusion as to the compatibility of this "system" can be made (if the susceptibility threshold of R_1 represents the susceptibility of R_1 in a valid manner). This is the <u>correct</u> method of calculating the compatibility of a system and is <u>exact</u> within the confidence of the generator, receptor and coupling models along with the confidence in the model of the intersystem effects.

The Boeing program approximates this correct solution in that the response of the receptor, $V_{R_1}(\omega)$, would consist of the sum of two received spectrums. One of which would be due only to G_1 with the parasitic nature of G_2 neglected and no consideration of the transmitter or intersystem effects. The other would be due only to G_2 with the parasitic nature of G_1 neglected and no consideration of the transmitter or intersystem effects. If one had "sufficient reason" to believe that the effects on R_1 due to the transmitter and intersystem sources was completely dominated by the effects due to G_1 and G_2, then within this assumption, one should add the spectrums due to G_1 and G_2 together. This would approximate the effect of G_1 and G_2 on R_1. It would still be an approximation since the Boeing program, for example, neglects the parasitic effect of the deactivated G_2 circuit when calculating the response of R_1 due to G_1. This

approximation of neglecting parasitic effects of other generator lines (as well as neglecting parasitic effects of other receptor lines) as is used by the Boeing program is apparently felt to yield worst case results. However, this may not necessarily be true in all cases. For example, resonance effects produced by including the parasitic nature of lines other than those of the generator receptor pair in the computation conceivably could produce receptor line responses of greater magnitude than those calculated on a generator receptor pair basis as in the Boeing program.

The Boeing program is an extension of other work in predicting cable coupling (and in certain programs, box radiation and reception).[33,37-39]

IV. THE McDONNELL-DOUGLAS PROGRAM[41]

This program was procured by AFAL for the EMC analysis of airborne communication-electronic systems and is currently in the final stages of development. Essentially the program determines through analytical models of antennas, wires and boxes as generators and receptors with associated coupling effects whether subsystem compliance to the appropriate MIL-STD's will insure compatibility of the system (within the confidence in the models of the generators, receptors and their coupling effects). Functional as well as nonfunctional sources are considered.

The program is modular in that each of the above coupling effects are calculated separately. For example, the wire-to-wire coupling is calculated as a separate routine; field-to-wire coupling is calculated separately, antenna-to-antenna is calculated separately, and box-to-box coupling is calculated separately.

The wire-to-wire coupling routine uses the lumped π equivalent circuit to represent the lines and their coupling effects. This routine differs from the Boeing program in that the entire wire generator, wire receptor network is solved simultaneously. In other words, the computation is not performed on a generator-receptor pair basis and the parasitic nature of other generator and receptor lines is included in determining the coupling between a generator-receptor pair. Nodal equations are written for the entire network under consideration and are solved by Gaussian elimination for the network node voltages. The per unit length parameters are derived for various shielding, grounding and twisting configurations. For frequencies where the wire length is greater than $1/20$ of a wavelength (for the frequency under consideration) the model of the

transmission line and its coupling effects consists of two π sections in cascade. In addition, only intrabundle coupling is considered due to computer storage limitations with the largest bundle consisting of approximately 60 wires with 4 shields, 40 wires with 20 shields or 20 wires all with shields. The calculation is performed in discrete frequency steps and the user can input the desired frequency range. The program has the capability to accept load and source impedances for generator and receptor wires as RLC networks, since it is essentially constructed as a general network analysis program. The source and susceptibility spectra are either the appropriate EMC specifications, measured spectral data (e.g., for known functional signals) or an internal generation of the magnitude of the Fourier Transform for the time domain output waveform of the generator. The program determines intermediate points between the specified discrete points in the spectrums by interpolating log-linearly between these points. The output of the program consists of the magnitude and phase of each source as well as the total received voltage due to all sources.

The antenna-to-antenna coupling model presently used is based on the results of [28]. This model is designed to predict coupling between antennas collocated on an aircraft. A transmitter antenna-receiver antenna pair is selected and checked for frequency coincidence. If frequency coincidence exists, then an EMI margin is determined at the <u>worst case</u> coincident frequency, f_ω, through

$$EM(f_\omega) = PR(f_\omega) - SR(f_\omega) \tag{4-1}$$

where PR is the power received from the transmitter at the worst case frequency and SR is the sensitivity threshold of the receiver at the worst case frequency. The power received from the transmitter, PR, is determined from

$$\begin{aligned} PR(f_\omega) = {} & P_T(f_\omega) + G_X(f_\omega) + G_R(f_\omega) - CL_X(f_\omega) \\ & - CL_R(f_\omega) - SF(f_\omega) + 20 \log_{10} \left| \frac{\lambda_\omega}{4\pi D} \right] \end{aligned} \tag{4-2}$$

where P_T, G_X, G_R, CL_X, CL_R, SF are the transmitter power, transmitting antenna gain, receiving antenna gain, transmitter and receiver cable losses and shading factor respectively. The shading factor, SF, reflects the attenuation caused by propagation around the fuselage and also includes a factor representing the diffraction effect of a wing separating the two antennas. The results are based on the work in [31] with similar results appearing in [29]. The last term in (4-2) is the Friis transmission equation where λ_ω is the wavelength of the chosen worst case frequency and D is the shortest distance around the fuselage between the two antennas.

The selection of the worst case frequency is performed in the following manner. One may expand (4-1) and (4-2) into

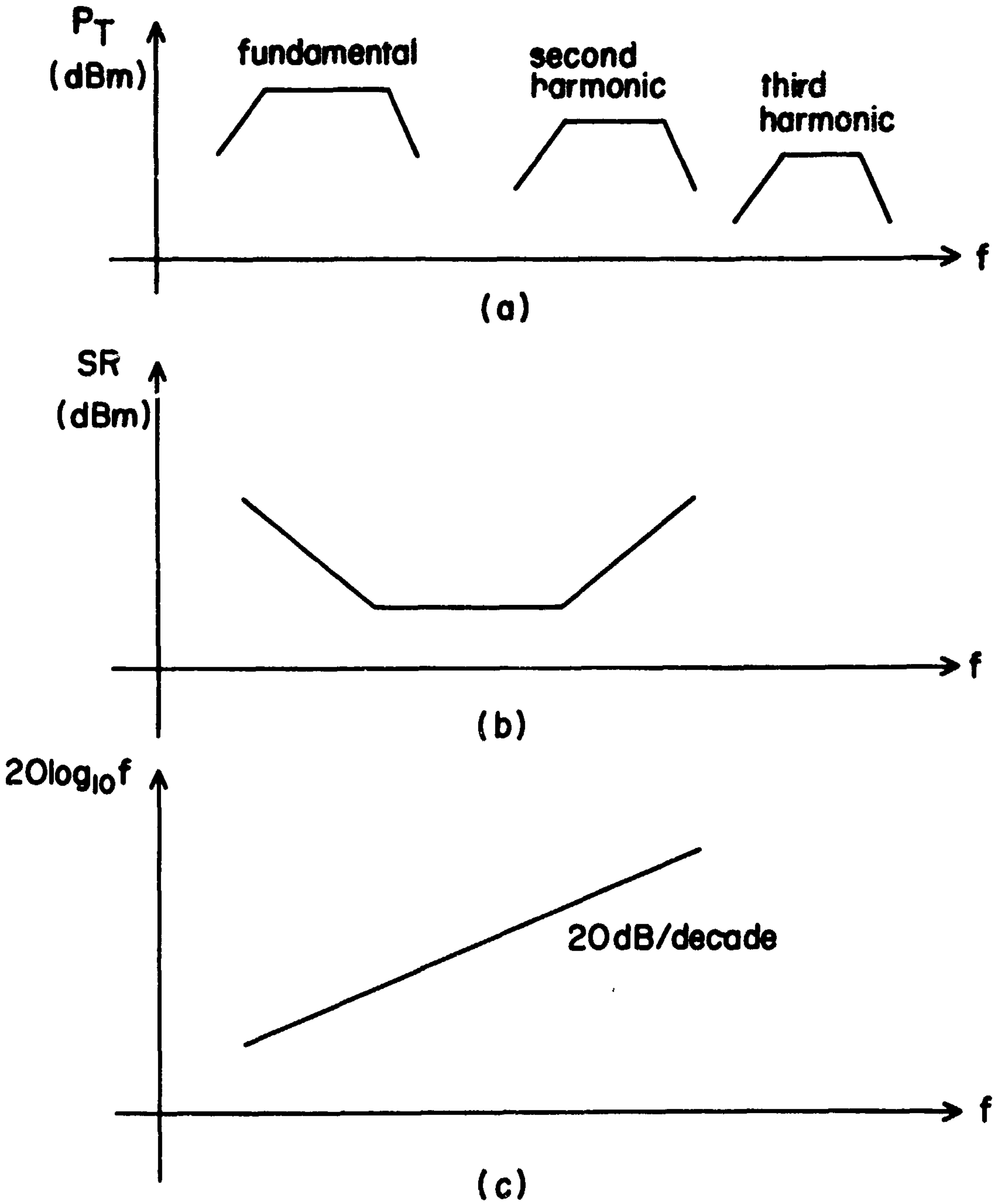
$$EM = P_T - 20 \log_{10} f - SR + K \tag{4-3}$$

where K is a constant for fixed G_X, G_R, CL_X, CL_R and D and the frequency dependence of SF is neglected. The transmitter power, P_T, is specified in Figure 36a, the selectivity of the receiver is specified in Figure 36b, and the term $20 \log_{10} f$ represents a 20 dB per decade slope. The term K in (4-3) is set to zero and the remaining three graphs are combined as in (4-3) to determine the worst case coupling frequency.

The gain functions are approximated by a two-level model; main beam gain and mean side lobe gain. Formulas are included for computing these gains in the absence of known data.

The program output consists of the maximum and minimum coincident frequencies with the selected worst case coincident frequency. The EMI margin for each transmitter-receiver pair is printed out.

The field-to-wire model is based on the results of [24] and translates an incident electromagnetic field into currents at either end of the line using the known line terminations and the distributed model for the line as described



P_T (dBm)
fundamental
second harmonic
third harmonic
f
(a)
SR (dBm)
f
(b)
$20\log_{10}f$
20dB/decade
f
(c)

FIGURE 36

in II. This model is primarily intended for use in predicting antenna-to-wire coupling through fuselage apertures such as gear doors, canopies, etc. In this useage, the antenna-to-antenna analysis program is used to determine the power incident upon the wire pair due to an on board transmitter. This incident power is converted to electric field strength assuming plane wave, free space propagation as described in II and a worst case field orientation is used for the field-to-wire coupling analysis. Allowances are made for shielding of the wires based on empirical data. The data output for this portion of the program consists of the EMI margin for the receptor wire and each transmitter. The total effect of all transmitters on the receptor wire is not obtained.

This program also has a routine for determining box-to-box coupling. This routine only considers power frequency magnetic box-to-box coupling at 400 Hz as well as the third harmonic, 1200 Hz. The transformer is considered as an N turn loop with current I. The expression for magnetic flux produced by this loop is easily obtained from elementary field theory (see [61]). The receptor is considered to be a power transformer feeding the input to a high gain audio amplifier. Consequently, the model of the receptor consists of an M turn, open-circuited coil and the induced voltage in this coil is compared to the susceptibility threshold of the amplifier in determining the EMI margin.

In this program, wire-to-wire, antenna-to-wire, antenna-to-antenna and box-to-box coupling are considered directly. However, one may also model box-to-antenna, box-to-wire, antenna-to-box with a rewriting of the program. However, there exists no model capability to determine wire-to-antenna and wire-to-box coupling.

In using this program, one must be careful to determine the total, or at least the "significant," energy incident upon a receptor from all sources. For example, the voltage appearing across the input terminals of a two terminal

receptor must be the <u>total</u> voltage due to coupling from <u>all</u> wires, cases and antennas (and intersystem sources if significant) if one wishes to relate the results to compatibility. If the coupling medium is assumed linear, then the program can still be executed properly in modular form to determine the energy incident on a receptor due to each source or groups of sources and these can then be added off-line to determine the total energy incident upon the receptor. Then the results can be related to compatibility.

As was discussed in III, the independent use of the wire-to-wire program implicitly assumes that the wire generators do not interact with box and antenna generators and wire receptors are assumed to not interact with box and antenna receptors. Also inherent in each of these modular subprograms (with the exception of the wire-to-wire routine) are the assumptions that the generators and receptors other than the pair being considered do not interact with each other or the pair under consideration.

V. THE TRW PROGRAM

We will discuss the philosophy of the program first and then consider the models used. This program represents a significantly new philosophy for the EMC community. As was discussed in I, the previous method of attack for considering intrasystem EMC consisted of applying somewhat rigid and rather general specifications to the individual subsystems (e.g., power and signal lines, antennas, etc.) comprising the total system. It was pointed out that these specifications were inadequate in many cases. The coupling between these subsystems has been considered generally only in an indirect way by adjusting the specification limits through waivers. This adjustment is in most cases based on engineering judgement. The difficulties in this regard are that incompatibilities may still surface during system test as a result of the human incapacity to predict precisely the highly complex coupling between the many subsystems. Also, the MIL-STD specifications do not easily admit trade-offs in the sense of required suppression, which would result in minimum system weight and complexity.

The TRW program develops subsystem specifications which are tailored to the individual systems. Obviously, each system will have different degrees of coupling between subsystems due to the different physical configurations and types of coupling paths between subsystems. Then if one determines analytically through a mathematical system model, which predicts this coupling, the signal levels coupled between subsystems, one should be able to arrive at specifications on the subsystems unique to each system which will achieve system compatibility with a minimum amount of unneeded suppression and restriction. The TRW approach is directed along these lines.

Of course, one may point out that there are many systems whose configuration and composition are not fixed. Also, many subsystems ("black boxes") are procured for general use in different types of systems. At first glance, this may seem to limit the program's applicability to fixed systems and subsystems designed for specific systems. Reference [2] contains valid arguments showing that this need not be necessarily true. The first argument against tailored specifications is that particular types of subsystems are procured for use in many different types of systems and these subsystems adhere to a fixed set of specifications instead of a set for each application. Tailoring specifications can still be used here since each particular system could be modeled and the composite minimum of the subsystem specifications generated for each system could be applied in the procurement of each class of subsystem which would guarantee their compatibility with these varied types of systems. On the other hand, if one does not wish to model each system and determine a set of subsystem specifications, then the actual subsystem test data could be utilized to determine if these subsystems would be able to integrate compatibly with each system considered.

The second argument against tailored subsystem specifications which are unique to each system is that in many cases the system environment is not fixed or describable, e.g., an overflying aircraft. In this case, it is impossible to determine if any subsystem limits would insure compatibility and the best "insurance" that one could obtain would be to suppress all undesired subsystem emissions and to reduce subsystem susceptibilities to the state-of-the-art. This would be the best we could do; however, this is clearly not cost effective. In this case, an approximation of the environment would be better than none at all.

Let us now discuss the TRW approach. A communication-electronic system consists of two types of sources of electromagnetic energy, desired or functional and undesired or nonfunctional. An example of a functional source is the fundamental frequency of a transmitter whereas the harmonics of this fundamental frequency are examples of nonfunctional sources. If we assume the coupling paths between sources and receptors are linear and time-invariant, then a frequency domain description of these paths, their inputs and outputs is equivalent to a time domain description through the Fourier Transform. For illustration, let us suppose that the system consists of p functional generators, q functional receptors and one nonfunctional generator and receptor each, represented by $G_1,\ldots,G_p$, $R_1,\ldots,R_q$, and G_N, R_N respectively as illustrated in Figure 37. Let us make the assumption that the receptors do not "load" the generators. Also assume that the generators do not interact and the receptors do not interact. With this assumption, we may represent the coupling paths as frequency domain transfer functions, $H_{ij}(\omega)$. For example, $H_{2N}(\omega)$ represents the transfer function between G_2 and R_N.

The outputs of all functional generators are frequency spectrums representing the Fourier Transform of the time domain outputs of these generators. The phase information is dropped and only the magnitudes of these Fourier Transforms are used yielding a worst case analysis. The susceptibility of the functional receptors is represented quantitatively by a single number AS_j. The susceptibility index AS_j is determined for a particular receptor by assuming that the receptor will respond to the <u>sum</u> of the incident frequency components, i.e.,

$$AS_j = \int_{f_1}^{f_2} P_j(f)\ \psi(f)\ df \qquad (5\text{-}1)$$

where f is frequency, $f_1 = 10\text{Hz}$, $f_2 = 10^{10}$ Hz, $P_j(f)$ is the "bandwidth" of the receptor port and $\psi(f)$ is the total spectrum incident at the receptor port.

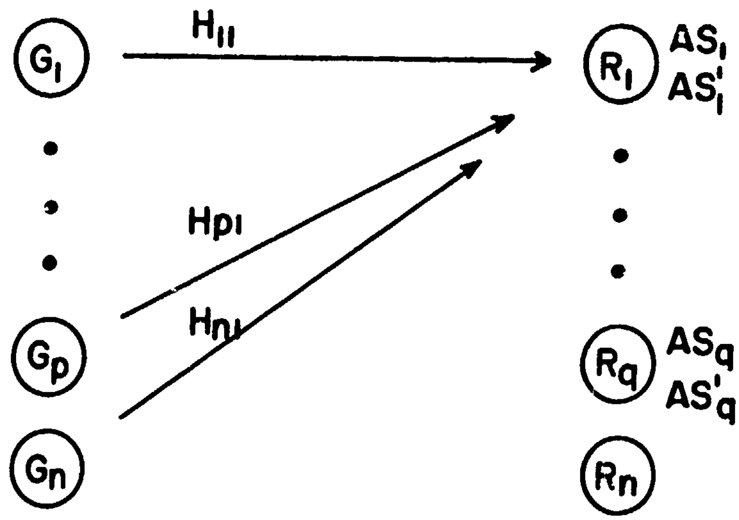


FIGURE 37

In practice, AS_j can be more easily determined from design considerations since the ports for functional receptors have well known properties, e.g., the input to a receiver, as opposed to nonfunctional receptor ports.

The program first determines functional compatibility between all functional generators and each functional receptor through a surplus margin AS_j' from

$$AS_j' = AS_j - \int_{f_1}^{f_2} P_j(f) \left\{ \sum_{i=1}^{p} H_{ij}(f)\ G_i(f) \right\} df \tag{5-2}$$

If $AS_j' < 0$ then the j-th functional receptor will be rendered susceptible due to functional sources. Then design modifications must be instituted since this functional receptor will certainly be interfered with when all sources (functional as well as nonfunctional) are in operation.

If all functional receptors are compatible with all functional generators then we will have q surplus margins AS_j' for $j=1,\ldots,q$ which are all greater than or equal to zero. The interference from the nonfunctional source G_N can be allotted to these surplus margins in the following manner. Compute for $j=1$

$$AS_1'' = AS_1' - \int_{f_1}^{f_2} P_1(f)\ H_{N1}(f)\ G_N(f)\ df \tag{5-3}$$

If AS_1'' is negative then reduce the spectrum of $G_N(f)$ (which is chosen on a worst case, arbitrary basis) so that (5-3) will now be nonnegative. Obviously there are many such reductions and spectrums of $G_N(f)$ which will yield a nonnegative AS_1''. Call this spectrum $G_N^{(1)}(f)$.

$$AS_2'' = AS_2' - \int_{f_1}^{f_2} P_2(f)\ H_{N2}(f)\ G_N(f)\ df \tag{5-4}$$

If AS_2'' is negative then reduce the spectrum of $G_N(f)$ so that AS_2'' is nonnegative. Call this resulting spectrum $G_N^{(2)}(f)$. Continue for each $j=1,\ldots,q$ to yield spectrums $G_N^{(1)}(f),\ldots,G_N^{(2)}(f)$. For each frequency (from a set of selected frequency points) between f_1 and f_2, determine the minimum of $G_N^{(1)}(f),\ldots,G_N^{(2)}(f)$.

Construct the spectrum $G_{N_T}(f)$ from these points. It then is clear that each functional receptor will be compatible with all of the functional generators and the single nonfunctional generator.

Now determine

$$R_N(f) = \left\{ \sum_{i=1}^{p} G_i(f)\ H_{iN}(f) \right\} + G_{NN}(f)\ H_{NN}(f) \tag{5-5}$$

To measure compatibility for this system then, we must test the nonfunctional receptor for susceptibility by applying $R_N(f)$ to this receptor in a laboratory situation.

For practical implementation, the transfer functions $H_{Nj}(f)$ and $H_{iN}(f)$ for $i=1,\ldots,p$ and $j=1,\ldots,q$ must be determined. To determine these, a "worst case" nonfunctional generator G_N and receptor R_N is assumed. The physical location of this generator and receptor is determined on a worst case basis for determining the above transfer functions.

The difficult off-line portions of the analysis concern the determination of a relevant and meaningful susceptibility index AS_j and also correct and meaningful worst case generator and receptor transfer functions above. These two areas are very crucial to the generation of a meaningful set of specifications (tailored to the specific system).

The question at this point is how the above analysis relates to developing a set of specifications to be applied to all nonfunctional generators in the same way MIL-STD-461 is used to insure system compatibility. Keep in mind that in the actual system test and deployment, each receptor (functional as well as nonfunctional) will be subject to the affects of all generators (functional as well as nonfunctional). Here we have developed a specification which would work for a system with only <u>one</u> nonfunctional generator. Obviously this is a highly unrealistic situation. However, if we divide the spectrum $G_{N_T}(f)$ by

the number of nonfunctional generators and apply the resulting spectrum to each nonfunctional generator, then we have insured compatibility if all nonfunctional generators are concentrated at the location of G_N and are the same type as G_N. If they are not, then we still will roughly add the effects of all nonfunctional generators on each of the functional receptors. This is of course an approximation since the transfer functions between each of these nonfunctional generators and each functional receptor will not be equal to $H_{N1},\ldots,H_{Nq}$. However, we are getting closer to uniquely tailored specifications. The TRW program in fact uses this method to develop the individual specifications for nonfunctional generators.

The individual subsystem susceptibility specifications are obtained in a similar manner by dividing $R_N(f)$, given by (5-53), by a factor representing the total number of nonfunctional receptors. The resulting spectrum is then applied in the laboratory to these nonfunctional receptors to determine if compliance to the specifications will relate to compatibility.

It should be pointed out that the program is quite flexible since as the system design progresses, test data on the nonfunctional generators and receptors will become available and can be brought in as data input with the result that the program can be used solely as an interference analysis program by classing all generators and receptors (functional and nonfunctional) as "functional" ones. Also, once the system has been modeled, then a rapid assessment of the impact of granting waivers as well as future system modifications can be assessed with a rapid turn-around time since the major portion of the off-line work has been done, i.e., the modeling of the system.

We should point out that one of the difficult problems in implementing this program will be the determination of receptor interference margins, AS_j,

and input bandwidths, $P_j(\omega)$. Note that we do not determine these parameters for nonfunctional receptors in the early design stages. These parameters are much easier to determine for functional receptors than for nonfunctional ones in the early design stages since the functional receptors have a great deal more information about their input ports and susceptibilities than nonfunctional ones (e.g., the input to a receiver as opposed to the power line input for a device).

Coupling Models:

All coupling determinations are performed on a generator-receptor pair basis. For example, the coupling between a wire-type generator and a wire-type receptor is performed by solving only the circuit consisting of this pair.

The wire-to-wire coupling analysis for a generator-receptor pair is determined in two parts. The mutual capacitive coupling between the wires is determined by neglecting mutual inductance. Next, the mutual inductance coupling is determined by neglecting the mutual capacitance coupling. The total coupling is then considered to be the sum of these coupling effects. Note that this is an approximation to the exact analysis of determining the total coupling by simultaneously solving for inductive and capacitive coupling. Also, as discussed above, only generator-receptor pair calculations are made yielding first order effects (i.e., the parasitic nature of other wires is neglected). In this computation the source data is a magnitude vs. frequency voltage or current source spectrum. Also, only the magnitude of the transfer function is used in the coupling calculation yielding a worst case result for this generator-receptor pair. A significant modeling difference exists between this wire-to-wire coupling computation and those of III and IV in that the distributive nature of the generator and receptor wires are not considered at all, i.e., neither the distributed model nor the lumped π model is used for the

wires themselves. Essentially this assumes that the voltages of the load and source output of the generator line are equal and the currents of the load and source output of the generator line are equal. Similar remarks apply to the receptor line. At present the load and source impedances are considered to be real valued resistances with a capability to include shunt capacitance. The program is not presently capable of accepting general complex impedances for source and load. Various shielding, grounding, and twisting configurations are included in the capability.

The antenna models are simply the specification of the E and H fields at the antenna site (or 1 meter away) and the replacement of the antenna with this plane wave. The input data needed is shown in Figure 38. So essentially the antenna is considered to be an isotropic point source radiator whose fields propagate outward uniformly in all directions and whose frequency dependence is as in Figure 38. One would have to specify f_1, f_2, and E_{min} and E_{max} off-line. The antenna H field is specified in the same way and both are specified one meter from the antenna.

For wire radiation (for example, wire-to-antenna coupling and wire-to-wire coupling at large distances) the H field at one meter from a wire parallel to a ground plane is determined in terms of the current on the wire using elementary field theory. Similarly, the E field at one meter from the wire is determined in terms of the wire voltage.

The field transfer models simply relate the above fields at one meter from the generator (antenna or wire) to the field at the receptor (antenna or wire) as in Figure 39. The field transfer represents a decay factor determined off line to relate the field F_2 at 2 to the field F_1 at 1; for example, $F_2 = (\frac{1}{r^n}) F_1$.

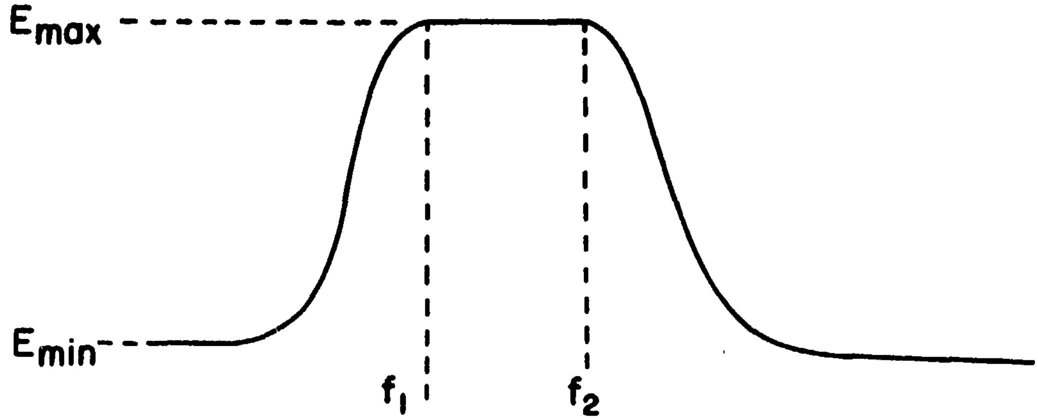


FIGURE 38

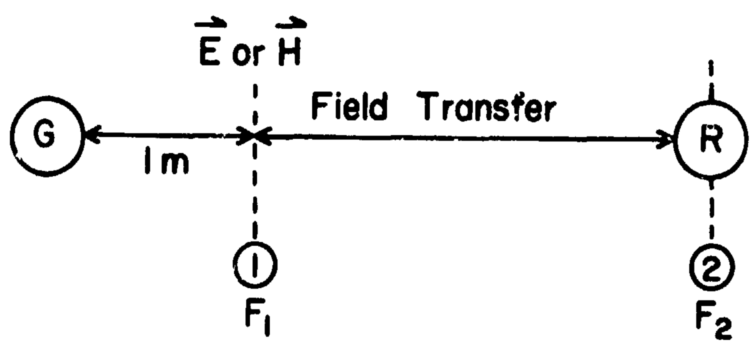


FIGURE 39

The field reception models relate the fields F_2 at the receptor to the voltage across the load for wire receptors and at the base of the antenna for antenna type receptors. The antenna reception essentially uses a frequency selection of the form in Figure 38.

In this program, the wire-to-wire coupling is calculated on a generator-receptor pair basis by adding capacitive and inductive transfer for each generator-receptor pair. For wire pairs separated over moderate to large distances, the E and H field coupling is computed for each generator-receptor pair. Also, the antenna-to-antenna, antenna-to-wire, antenna-to-box, box-to-wire, box-to-box, box-to-antenna, wire-to-box, and wire-to-antenna coupling is calculated on a generator-receptor pair basis.

VI. THE SACHS-FREEMAN PROGRAM[35]

This work was procurred by the Electronic Sy.,tems Division (ESD) of the U. S. Air Force. We will discuss the portion of the program dealing with the intrasystem compatibility investigation via compliance to MIL-STD-461.

This portion of the program is focused on determining whether subsystem compliance to MIL-STD-461 will insure overall system compatibility. The investigation consists of three parts.

The Class I investigation determines box-to-wire and wire-to-wire coupling and uses the wire-to-wire and field-to-wire coupling models developed by TRW and described in V. This analysis considers wire-to-wire and box-to-wire generator-receptor pairs.

For wire-to-wire coupling, the routine selects the applicable conducted narrowband limit (from CE01, CE02, CE03 or CE04) at the frequency of the receptor. This conducted limit represents a maximum allowed value of current on the emitter line and is coupled via inductive transfer to the receptor line resulting in a voltage across the input to the receptor, V_{NCI}. This voltage is a single numerical value and not a frequency dependent spectrum since only the receptor frequency was considered. Next, the routine divides the value of the above narrow band conducted limit (obtained by evaluating the appropriate CE limit at the receptor frequency) by the emitter line impedance resulting in a voltage on the emitter line. It is not clear which emitter line impedance is considered; for example, the emitter line load impedance or the impedance seen by the source. This emitter line voltage is coupled to the receptor line through capacitive transfer resulting in a voltage across the input to the receptor, V_{NCC}. Again, V_{NCC} is a single number and not a frequency dependent spectrum since only the receptor frequency was considered.

The above calculations are repeated using the applicable broadband limits in MIL-STD-461 resulting in V_{BCI} and V_{BCC}, the inductive and capacitive coupled broadband limits as coupled to the receptor line respectively. The broadband limits are multiplied by the bandwidth of the receptor circuit in making these calculations.

The values of the radiated emission limits (RE01, RE02) are determined apparently at the frequency of the receptor. Each wire coupling pair separated at moderate distances (those not considered above) and each case-to-wire coupling pair are now considered by transferring the values of RE01 or RE02 selected at the receptor frequency to the receptor wire through the TRW field-to-wire coupling models resulting in voltages at the receptor input port V_{NE}, V_{NH} due to the narrowband limits and V_{BH} due to the narrowband H field emission limit, RE01.

The above signal levels for narrowband and broadband interference are added together to produce

$$V_B = V_{BCI} + V_{BCC} + V_{BH} \qquad (6\text{-}1)$$

and

$$V_N = V_{NCI} + V_{NCC} + V_{NE} + V_{NH} \qquad (6\text{-}2)$$

where V_B is the broadband coupled interference and V_N is the narrowband coupled interference each of which is due to wire and box radiators only and are given in volts. Each value is a single real number.

At this point, one must specify a threshold voltage AS at which the receptor malfunctions. This quantity is a single number given in volts. The susceptibility of each wire receptor to all wire and box emitters only is determined by summing the total narrowband received voltage at the receptor input, V_{N_T}, and the total broadband received voltage at the receptor input, V_{B_T}. The susceptibility of this wire receptor is determined from

$$ACTN = AS(CM) - V_{N_T} \qquad (6\text{-}3)$$

and

$$ACTB = AS(CM) - V_{B_T} \tag{6-4}$$

where CM is a safety margin. If ACTN or ACTB is negative, then the wire receptor is said to be susceptible.

The second class considers antenna-to-wire coupling. The electric field incident upon a wire receptor due to a transmitting antenna is calculated with the EPM-1 path loss model[39] and is then converted to receptor input wire voltage with the TRW field-to-wire reception model as follows. The power received by a hypothetical isotropic antenna in free space at the receptor wire location is calculated through

$$P_R = 10 \log P_T + G_T + LPROP + 30 \tag{6-5}$$

where P_R is the received power of this isotropic antenna, P_T is the power of the transmitter at its midband operating frequency, G_T is the gain of the antenna in the direction of the wire and LPROP is the value of the free space path loss between the antenna and the wire receptor. This received power is converted to electric field strength, E, using the assumption of uniform plane wave propagation as described in II. The TRW E field-to-wire transfer model is then used to obtain the resulting voltage across the receptor terminals due to the effect of this transmitter. These values are then compared to the susceptibility threshold number, AS, for the wire receptor. Apparently the effect of each transmitter on the receptor is compared to AS separately and if any one effect exceeds AS then the transmitter-receptor wire pair is said to be incompatible or at least a potential problem area exists.

Class three represents wire, box and transmitting antenna-to-box coupling. For all emitter-to-box combinations, the program calculates the E and H field signal levels at the receptor case and tabulates these. It is not clear whether this tabulation is the sum of the effects due to each emitter or whether the

effects are tabulated individually. Furthermore, it is not clear whether these tabulations are levels of received signals versus frequency or at a single frequency; if so, what frequency is used. No further calculations are made and apparently one would be expected to perform laboratory tests on the box of interest by simulating these tabulated field strength(s).

Some discussion of this program is in order. In the calculation of wire-to-wire and box-to-wire coupling, the value of the applicable limit of MIL-STD-461 is selected at the wire receptor's frequency. It is not clear that all wire type receptors are susceptible at only one frequency. For example, a device is not necessarily susceptible to signals entering the power line input at only the power supply frequency. To perform a proper analysis in this regard, one should determine the coupled signal levels at either all frequencies or only those for which the device is susceptible at the wire input port (if such a determination can be made).

Finally, the modular useage of this program will point out only certain areas of susceptibility. For example, if one determines a wire type receptor's port voltage due to all box and all other wire couplings and if this value exceeds AS (the susceptibility value of the receptor) then one can surely state that this receptor will be interfered with. This is, of course, completely dependent upon a meaningful susceptibility value, AS. However, if this value does not exceed AS, then can one say that this wire receptor will operate compatibly with the system? Obviously this statement cannot be made with confidence. This is due to two considerations. First of all, this receptor may be susceptible to signals other than the single frequency chosen for the analysis. We have not considered incident signals at other frequencies. The second reason is a result of the fact that we have not added in the antenna and box coupled effects. Even if there exist no antennas within the system whose fundamental

frequency is the same as the chosen wire receptor frequency, incompatibilities may result since this wire receptor may be susceptible to frequencies other than the one designated as the "receptor frequency" and in particular one of the transmitter fundamental frequencies. Also, one should consider not only the fundamental frequency of the transmitter but also the spurious responses.

VII. CONCLUSION

In conclusion, we may review our discussions with a brief comparison of the philosophies and methods of attack in each program.

First of all, we might point out that none of the above programs consider intersystem effects. Of course, the modeling and inclusion of intersystem effects can be a difficult problem, yet they will effect each system in an actual deployment and should be considered.

As for philosophies, basically the Boeing, McDonnell-Douglas, and Sachs-Freeman programs consist of an analytical capability to determine the electromagnetic energy coupled between subsystems. This capability is used to determine whether or not compliance to the subsystem specifications, e.g., MIL-STD-461, will ensure system compatibility. This determination is crucially dependent upon having a meaningful quantitative measure of the susceptibility of a device. The Boeing and McDonnell-Douglas programs use a frequency dependent susceptibility threshold as in the MIL-STD's which, as was pointed out in the introduction, needs justification since devices basically respond to total incident energy. The use of a broadband susceptibility criterion provides only a rough measure of this. The Sachs-Freeman program appears to consider a device susceptible at only one frequency which seems rather unrealistic. The TRW program on the other hand considers a device susceptible to a measure of total incident energy since the effect of an incident spectrum is summed.

As for models of the coupling paths between subsystems, all programs are in universal agreement as to the use of a frequency domain description of the coupling paths as opposed to a time domain description. As discussed in the introduction, this seems to be a valid approach. The wire coupling models use

lumped approximations to the exact distributive calculation. Apparently this approximation is considered to outweigh the suspected increase in computation time with the more exact distributive model. These approximations, however, only are valid for frequencies where the circuit dimensions are much less than a wavelength and it has not been conclusively demonstrated that the distributive calculation necessitates the large increase in computational time over the lumped approximations as seems to be expected.

All programs essentially perform the calculations of transferred energy on an emitter-receptor pair basis and assume that other generators and receptors will not affect the pair being considered in the calculations. This obviously reduces the required computation time substantially since the size of the investigated circuits (for wire-to-wire coupling) generally requires the minimum computational effort. The one exception to this generator-receptor pair calculation is the McDonnell-Douglas wire-to-wire coupling program.

The antenna coupling calculations also assume no interaction between emitters and between receptors. For example, the mutual coupling between transmitting antennas is not considered. Also no computations of intermodulation, crossmodulation or spurious responses are included for receivers attached to antennas. The propagation models use the free space path loss equations described in II which rely heavily upon obtaining accurate antenna gain factors (in the direction of transmission). Furthermore, these gain factors are generally assumed independent of frequency which seems to be an unrealistic assumption.

Box radiation, reception and coupling has been minimally considered apparently for the reasons given in II.

The most important point to be brought to the attention of users of these types of programs is the following. In the actual system deployment, each

subsystem will be effected to some degree by all sources (wire, case, and antenna coupled effects as well as intersystem effects). If one determines the effects on a receptor due to wire type sources only and obtains the result that the receptor will be interfered with, then it is valid to conclude that this receptor will be rendered susceptible in a system test. If, however, the receptor is found to be not susceptible to only wire sources, then one cannot state that the receptor will not be interfered with during system test since the effects due to other sources (intersystem, case, and antenna) have not been added in. Considering partial effects of sources on a receptor will not necessarily yield the complete system compatibility picture!

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